Proceedings of the I·R·E

A Journal of the Theory, Practice, and Applications of Electronics and Electrical Communication

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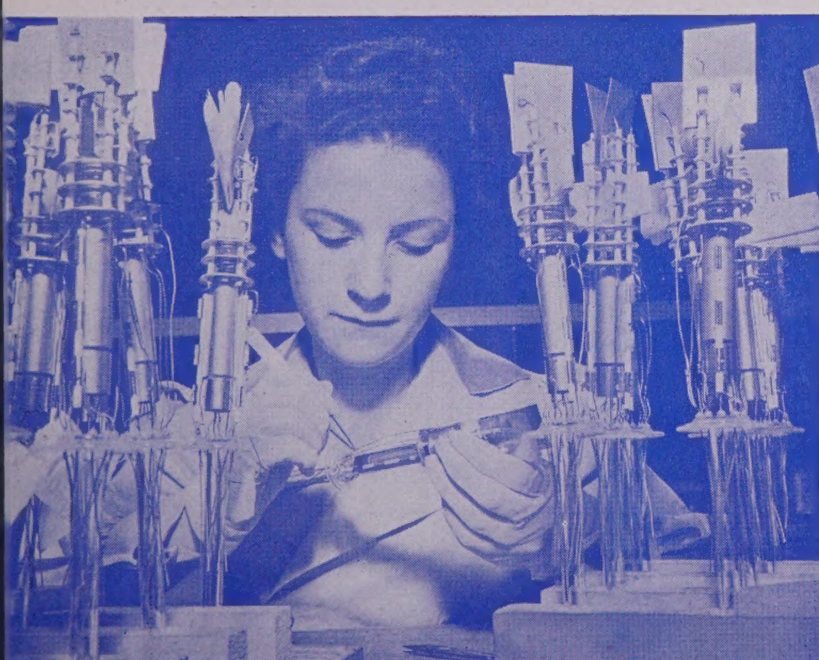

National Film Board

GREAT-GRANDCHILD OF THE DIODE
Cathode-ray tube mounts checked in air-conditioned room.

SEPTEMBER, 1945
Volume 33 Number 9

The Institute of Radio Engineers

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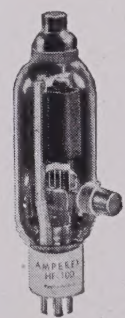

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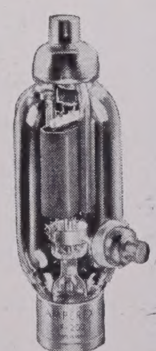

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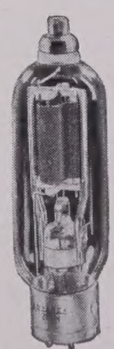

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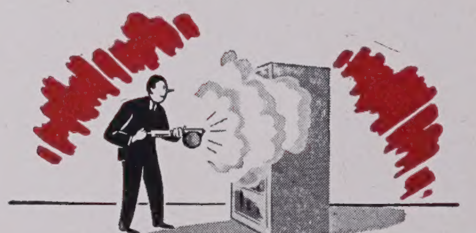

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Proceedings of the I·R·E

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Responsibility for the contents of papers published in the PROCEEDINGS rests upon the authors. Statements made in papers are not binding on the Institute or its members.

The Institute of Radio Engineers

INCORPORATED

SECTIONS

Sections	Next Meeting	Chairman	Secretary	Address of Secretary
Atlanta	September 28	R. A. Holbrook	F. W. Fisher	Georgia School of Technology, Atlanta, Ga.
Baltimore	—	R. N. Harmon	H. L. Spencer	714 S. Beechfield Ave., Baltimore, Md.
Boston	—	C. C. Harris	Corwin Crosby	16 Chauncy St., Cambridge, Mass.
Buenos Aires (Argentina)	—	A. DiMarco	H. Krahenbuhl	Transradio Internacional, San Martin 379, Buenos Aires, Argentina.
Buffalo-Niagara	September 19	J. M. Van Baalen	H. W. Staderman	264 Loring Ave., Buffalo, N.Y.
Cedar Rapids	September 20	F. M. Davis	J. A. Green	Collins Radio Co., 855–35 St., N.E., Cedar Rapids, Iowa.
Chicago	September 21	Cullen Moore	L. E. Packard	General Radio Co., 920 S. Michigan Ave. Chicago 5, Ill.
Cincinnati	—	L. M. Clement	J. F. Jordan	The Baldwin Co., 1801 Gilbert Ave., Cincinnati 2, Ohio.
Cleveland	September 25	H. B. Okeson	A. J. Kres	16911 Valleyview Ave., Cleveland 11, Ohio.
Connecticut Valley	September 20	H. W. Sundius	L. A. Reilly	989 Roosevelt Ave., Springfield, Mass.
Dallas-Fort Worth	—	J. D. Mathis	B. B. Honeycutt	9025 Roanoak, Dallas 18, Texas
Dayton	September 20	L. B. Hallman	Joseph General	411 E. Bruce Ave., Dayton 5, Ohio.
Detroit	September 21	L. H. Larime	R. R. Barnes	1411 Harvard Ave., Berkley, Mich.
Emporium	—	W. A. Dickinson	H. E. Ackman	West Creek, R. D. 2, Emporium, Pa.
Indianapolis	—	H. I. Metz	E. E. Alden	4225 Guilford Ave., Indianapolis, Ind.
Kansas City	—	R. N. White	Mrs. G. L. Curtis	6003 El Monte, Mission, Kansas.
London (Canada)	—	B. S. Graham	C. H. Langford	246 Dundas St., London, Ont., Canada.
Los Angeles	September 18	R. C. Moody	R. G. Denechaud	Blue Network Co., 6285 Sunset Blvd., Hollywood 28, Calif.
Montreal (Canada)	—	L. A. W. East	R. R. Desaulniers	Canadian Marconi Co., 2440 Trent Rd., Town of Mt. Royal, Que., Can.
New York	—	G. B. Hoadley	J. T. Cimorelli	RCA Manufacturing Co., 415 S. Fifth St., Harrison, N.J.
Ottawa (Canada)	—	W. A. Steel	L. F. Millett	33 Regent St., Ottawa, Ont., Can.
Philadelphia	October 4	D. B. Smith	P. M. Craig	Philco Corp., Philadelphia 34, Pa.
Pittsburgh	September 10	J. A. Hutcheson	C. W. Gilbert	Box 2038, Pittsburgh 30, Pa.
Portland	—	Kenneth Johnson	C. W. Lund	Route 4, Box 858, Portland, Ore.
Rochester	September 20	G. R. Town	A. E. Newlon	Stromberg-Carlson Co., Rochester 3, N.Y.
St. Louis	—	B. B. Miller	N. J. Zehr	KFUO, 801 DeMun Ave., St. Louis, Mo.
San Diego	—	F. A. Everest	Clyde Tirrell	U.S. Navy Radio and Sound Laboratory, San Diego 52, Calif.
San Francisco	—	David Packard	William Barclay	Stanford University, Calif.
Seattle	September 13	G. L. Hoard	K. A. Moore	5102 Findlay St., Seattle 8, Wash.
Toronto (Canada)	—	F. H. R. Pounsett	Alexander Bow	Copper Wire Products, Ltd., 137 Roncesvalles Ave., Toronto, Ont., Canada.
Twin Cities	—	Ross Hilker	Merle Ludwig	Minneapolis-Honeywell Regulator Co., Minneapolis, Minn.
Washington	October 8	H. A. Burroughs	L. C. Smeby	4801 Connecticut Ave., N.W., Washington, D. C.
Williamsport	October 3	L. E. West	F. L. Burroughs	2030 Reed St., Williamsport 34, Pa.

SUBSECTIONS

Subsections	Next Meeting	Chairman	Secretary	Address of Secretary
Columbus	—	E. C. Jordan	Warren Bauer	376 Crestview Rd., Columbus 2, Ohio.
Fort Wayne			—	
Milwaukee	—	P. B. Laeser	E. L. Cordes	3304 N. Oakland Ave., Milwaukee, Wis.
Monmouth	—	L. J. Giacoletto	C. D. Samuelson	5 Russel Ave., Ft. Monmouth, N. J.
Princeton	—	W. C. Johnson	J. G. Barry	Princeton University, Princeton, N. J.
South Bend	—	H. E. Ellithorn	J. E. Willson	1002 S. Lombardy Dr., South Bend, Ind.

The PROCEEDINGS enjoys the privilege of presenting to its readers the viewpoints of the I.R.E. Section Chairmen, in the form in which they are received. Such expressions cannot but be constructively helpful to the I.R.E. membership and its administrative officers alike. Accordingly there follows a stimulating guest editorial from the Chairman of the Emporium, Pennsylvania, Section.

The Editor

The Role of the Sections in the I.R.E. Program

WILLIAM A. DICKINSON

I want to analyze briefly the functions of the Sections, and to affirm my belief in their importance to the development of the I.R.E.

The Section is first the local agent of the Institute. As such, its primary duty is the holding of technical meetings of interest and benefit to its members. I know that the Section program committees are doing an excellent job under present limitations. Aids to the Sections, now being planned by the Institute, in improving and expanding their technical programs, should reap substantial rewards in interest and new membership.

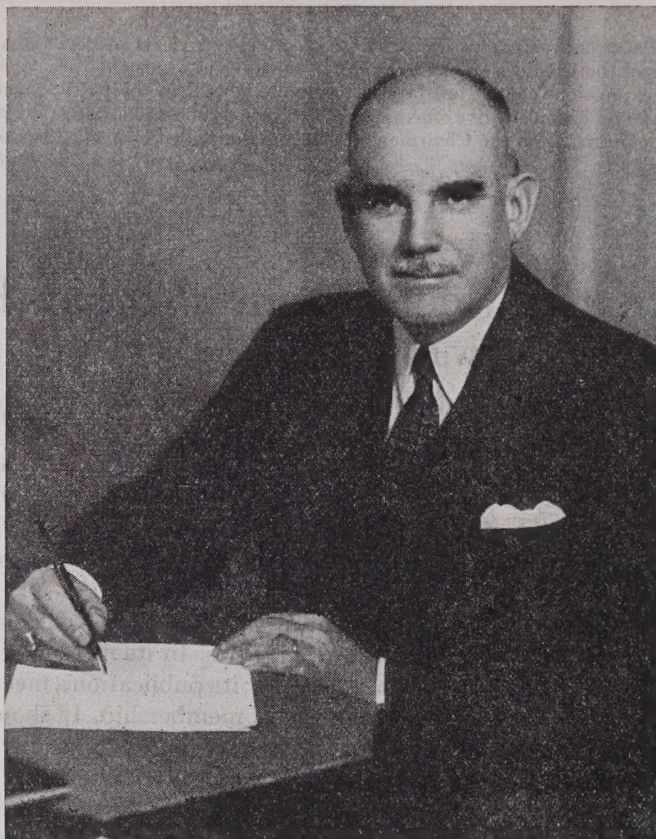
The Section has an obligation to keep its members, and local firms engaged in the electronics business, informed of and interested in Institute activities. In its meetings, bulletins, and news releases it should advertise the services of the Institute: its publications, meetings, and technical-committee work, and should stress the advantages of membership. It should actively support Institute projects—such as the Building-Fund Campaign—and will usually find its efforts well repaid in the interest resulting from direct participation by the members.

The Section should provide means for expressing to the Institute the ideas, wishes, and interests of its members. This duty is too often overlooked. Actions of the Institute officers and Board which draw criticism from the membership may often result from a complete lack of expression of opinion from the members, either personally or through the Sections. If the Sections were more aggressive is encouraging expression of opinion, the Institute could be better guided to the satisfaction of all of us.

The Section is also responsible for keeping the local Institute records and for maintaining a healthy condition of membership in its area. Probably every Section has its problems with local rivalry and with groups which could become useful members, but who choose to enjoy a purely passive membership. A working Personnel Committee can be most useful to the members, the Section, and the Institute, with suggestions for appointments and upgrading. I believe that the efficiency of the whole Section executive committee could be much improved by a visit from a member of the Executive Secretary's staff early in his term of office. Many Section officers have little preparation or aid for their duties except the use of the Section Manual.

Aside from its functions as an agent, the Section is the natural medium for acquainting members with each other's work and technical interests and problems. Most Sections can stimulate interest by encouraging their own members to present technical papers or brief reports on subjects of which they have special knowledge.

These, in general, are the Sections' obligations and opportunities. The burden of performance usually falls on the Section Executive Committee. This group consists of volunteer workers, and to do its job well, needs the assistance of the members and of the Institute staff. The Institute knows that strong Sections mean a stronger I.R.E., and most members realize that active participation returns them maximum membership benefits. Let's promote our Section programs.



Blackstone

George W. Bailey

Executive Secretary, The Institute of Radio Engineers

George W. Bailey was born in Quincy, Massachusetts, May 14, 1887. He received the A.B. degree from Harvard College in 1907, and is treasurer of his class.

He first engaged for fifteen years in the manufacturing of shoes in New England, and then became active for another fifteen years in rubber manufacturing.

He has long been well known as an amateur radio operator, and his station call W1KH is recognized the world over. Since 1940 he has held the high offices of president of the American Radio Relay League, and president of the International Amateur Radio Union. His standing has been further recognized by the award to him of the Marconi Medal of Service of the Veteran Wireless Operators Association, of which organization he is an honorary member.

In May, 1941, he was invited to go to Washington by the National Research Council, in order to assist the Signal Corps in recruiting candidates, with scientific training, suitable for the Electronics Training Group. Later he became chief of the Scientific Personnel Office, of the Office of Scientific Research and Development, under Dr. Vannevar Bush, director.

He enjoys an exceptionally wide acquaintance among scientists and professional engineers, and has an unusually intimate knowledge of the important part they have played in the successful prosecution of the war.

Mr. Bailey recently was appointed Executive Secretary of The Institute of Radio Engineers. He will also continue to work with Dr. Bush for an indefinite period, in the meantime devoting part of his time to his duties with the Institute.

In 1938 he became an Associate Member of the I.R.E. He is the Chairman of Committee 1, and a member of Committees 12 and 13 of the War Communications Board, and a member of the Administrative Committee of the Radio Technical Planning Board. He is Chief Radio Aide of the War Emergency Radio Service of the District of Columbia. He is a member of the Cosmos Club of Washington, D. C., and the Harvard Clubs of Washington, New York, and Boston.

The Organization of Research in the Radio Industry After the War*

W. RUPERT MACLAURIN†

Summary—**One of the serious weaknesses in the organization of research in the radio industry between the two world wars was the absence of adequate research or incentives to research, on the part of RCA licensees. Another weakness was the tendency to bureaucracy in the large industrial laboratories.**

Our major postwar scientific need is for more men who combine a thorough training in their subjects with creative originality. Such men can be produced, but industry must give further thought to the conditions which lead to the most effective flowering of group and individual scientific productivity. There has been a tendency in the radio and other industries to burn up our young research men and exploit their youthful energy.

There were very few companies in the radio industry prior to the war that had acquired the research spirit. Most of them were essentially imitative. The major rewards in these companies went to manufacturing positions, and the research men were given very little freedom and were placed under considerable pressure for short-range results. Where significant research contributions were made, adequate recognition was not given. This pattern should be changed, but it cannot be done merely by giving lip service to research.

Another vital task is the protection of research personnel during business depressions. If we go through a period of difficult postwar readjustment, there is a serious danger that the research interest which is now developing in the medium-sized and small companies will not survive. This would be very unfortunate, because the health of the radio industry will depend on the number of vigorous and vital independent research organizations that it supports after the war.

AT THE OUTSET I should like to explain how a professor of industrial economics happens to be concerning himself with the organization of research in the radio industry. About four years ago, a new program of studies was initiated at the Massachusetts Institute of Technology on the economics of technological change, under an exploratory grant from the Rockefeller Foundation. Our purpose has been to undertake studies in a number of different industries in order to assess the economic-environmental conditions which have been conducive to effective technical progress. Such studies, we hope, will prove helpful in indicating some of the postwar research goals for which we should strive and some of the methods that might be utilized for reaching those objectives. In this venture the economist needs help and encouragement from the engineer. I hope, therefore, that we may receive ideas and suggestions from the radio engineers for carrying out these studies, especially the study of the radio industry in which I have been actively engaged for some time.

Research objectives in the radio industry after the war should be part of the more general problem of research goals for American industry. What do we want from such research? In the first place, we want steady progress to higher standards of living; and research has a very vital role to play in achieving that objective. We also want to provide adequate outlets for new investment in American enterprise. A mature economy, with a high level of individual and corporate savings, needs increasing outlets for investment to continue its development. If, as is often suggested, our American frontiers for the balance of the twentieth century are likely to be scientific, we must spend a considerably higher proportion of our national income on research. We must also lay greater emphasis on pure research; for it is from pure research that the basic ideas for new industries and important new products spring. I believe, therefore, that the scientists and research workers in the United States can look forward to increasingly important roles in American economic life in the years ahead. But it will take serious thought by industry, the universities, and the government to achieve a rate of scientific progress that is commensurate with our national needs and potentialities.

The term "research" can be used to cover production engineering, advanced engineering development, and pure research. For the purposes of discussion here, I am concerned exclusively with the last two functions. I am also primarily concerned with original advanced engineering development rather than with imitative development.

The radio industry stems originally from pure research which was carried on by a handful of university professors in the last half of the nineteenth century. It is to Faraday, Maxwell, Hertz, J. J. Thomson, Branly, Popoff, Oliver Lodge, and Wehnelt that we must look for the origins of the industry. These men worked either in universities or in institutions of a comparable character, such as the Royal Institution in London, where Faraday conducted his vital research program.

From 1900 to 1920 the fundamental discoveries that had been made before the turn of the century were first reduced to practice. The work of Marconi, de Forest, Fessenden, and Armstrong can be characterized, I believe, as advanced engineering development of the best type. So, also, were the contributions of Arnold, Nicholson, and others of the Bell Telephone Laboratories; and of Langmuir, Alexanderson, and Hull, of the General Electric Company. The dividing line in some of these cases between what we should characterize as pure research and advanced development is difficult to draw. The contributions of Langmuir in particular lay at least equally in pure research.

What are the high spots of the period from 1920 to 1940? It is to the General Electric Company, the Bell Telephone Laboratories, and the Westinghouse Elec-

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† Massachusetts Institute of Technology, Cambridge, Mass.

tric and Manufacturing Company that we owe the very important developments in power tubes. The General Electric Company produced the screen-grid tube. The Westinghouse Company and the Radio Corporation of America brought out the iconoscope, with its revolutionary impact on television. Farnsworth, working alone, produced the image dissector; and Armstrong, as a college professor and independent inventor, developed frequency modulation. There were, of course, hundreds of other significant improvements that took place in radio technique, but these were some of the brightest spots.

Although this was a period in which substantial contributions were made, I believe that there were serious weaknesses in the organization of research in the radio industry between the two world wars. One such weakness was the absence of adequate research or of "incentives to research" on the part of RCA licensees during this period. Another weakness was the tendency to bureaucracy in the big industrial laboratories. Their contributions were important, but there was too high a proportion of directed projects. In physics, for example, there is evidence that the large industrial laboratories after World War I did not attract the best young physicists and give them freedom to develop their science as they saw fit. University research in physics and electrical engineering was inadequate in volume and freedom during this period. By and large, many of our better university departments, which were capable of making research contributions, were harassed by heavy teaching loads and seriously restricted by financial limitations. There were too many short-range projects undertaken; and there were too many engineering professors who were forced to supplement their incomes by short-term consulting relationships with industry. Such consulting arrangements had certain advantages in keeping close contact between universities and industry, but in many cases they were carried to the point of inhibiting sustained research effort.

What can we expect in the electronics industry from 1945 to 1965? Fortunately, many more companies have had a taste of research during the war, and have initiated programs which they will try to continue. There is also a significant movement toward integration in the radio industry which is resulting in the emergence of large, integrated units, each capable of conducting substantial research. The period of novitiate of the Radio Corporation of America is over, and, separated from its parents, the child may well expect vigorous competition from both the General Electric Company, its father, and the Westinghouse Company, its mother, who are getting to be more independent of each other than they ever were before. Philco has absorbed National Union; Sylvania has taken on Colonial; Bendix, so far an unknown quantity, is expected to enter both the set and tube industry after the war as a large and important factor with which to be reckoned. These companies alone constitute substantial, integrated concerns where there was only one during most of the period from 1920 to 1940. Such a situation is considerably more healthy.

The subsequent question, however, is whether organizations like Philco, Sylvania, Crosley, Emerson, Galvin, and Zenith will build up important research organizations after the war; and will the smaller companies make comparable contributions relative to their size? If every company in the industry supported a first-class research organization it certainly would not pay as well as in the past. However, I think this very unlikely. The majority of small companies cannot support vital research organizations. Moreover, there will always be imitator firms, firms which keep their research budgets purposely low and make profits from sales and production skill and the speed with which they are able to follow somebody else's lead. Such companies perform a useful function in getting new products introduced rapidly. It would be unfortunate if all companies in an industry were imitator firms, but we need some such concerns and we are certainly going to have them.

My own belief is that research in electronics will concontinue to pay, although not the dividends, probably, that the General Electric Company obtained from its original investment in Whitney, Coolidge, and Langmuir. Those returns were enormous, and they were thoroughly justified from the social point of view. Yet, while the return is not likely to be as large as this, the possibilities in electronics are still sufficiently great that ample dividends will be obtained from intelligent investment in research. It is up to the engineers to persuade their managements that research pays and will continue to pay. More important, they will need to persuade their managements what it means. There has been too much lip service to research in this and other industries, a confusion of imitative development work with the function of creating important new ways of doing things. There is a serious danger that the current popular interest in research which has infected management as well as the public at large will result in the mushrooming of second-rate research groups throughout the country. If research is to mean anything at all, it must be well done. Otherwise it is futile and a waste of time.

The smaller industrial companies interested in moving out of an essentially subservient and imitative position have a particularly difficult task. They should offer considerably greater rewards than they have in the past for men who make original contributions. They also should encourage men to plan a research career and stay with it. Too often the major rewards go to manufacturing positions or to administrative positions in research. This needs to be supplemented by rewards for individual brilliance. Research has moved in the direction of co-operative teams, but this should not be the only pattern. For a man of high originality who is not able to work well in a co-operative team, special efforts

ought to be made to find an important place in the company's organization. The smaller companies by and large have not faced this problem squarely. If the rewards in a company in terms of salary, prestige, and security go to the administrator, ambitious young men will be continually pushing in that direction, regardless of whether they have any administrative skills.

Those who are responsible for directing research in the organizations which are beginning to get a taste for research will need to convince their managements of the necessity of the long-range approach to research accomplishment. They will need to demand that they have some men of high quality on their staffs who are not expected to deliver in a short period, who should be relatively free within the framework of the company's interests to study and develop as they see fit. Such men, if they wish to, should be encouraged and financed by their companies to do postgraduate work in a university, and they should be encouraged to publish the results of their work. The publication of significant results is important as an intellectual discipline and as a method of gaining the interest and respect of one's professional colleagues, without which interchange of ideas does not come freely.

I believe that, if a company's field of interest is defined, one can rely on the intelligence of a good man to develop a fresh approach to his company's problems. Management, however, must expect that there will be some failures in research. If a company is not willing to finance and encourage research with the degree of freedom that I have suggested, it has not acquired the research spirit; and I believe that the further spread of the research spirit is not only very important from the standpoint of our national interests but will continue to pay dividends to those industrial concerns that support it.

Much has been said about the shortage of highly trained and original scientific personnel. There will be a shortage after the war, but if industry will offer interesting and important career opportunities for such men, they can be created. To do so may mean reaching down into the high-school system to encourage youngsters of scientific talent. There is a growing realization of the importance of doing this, and industry must play its part in the process. The work of the Westinghouse Company along these lines deserves special commendation.

Industry must give further thought, however, to the conditions which lead to the most effective flowering of group and individual scientific productivity. This is perhaps an area of inquiry in which the universities can be of service. Our industrial relations section at Massachusetts Institute of Technology is expanding its work in the field of psychology; and one of our projects is to make some studies of scientific productivity in industry and in universities. We should be very glad to know of companies which would be interested in cooperating with us in studies of this nature. Preliminary inquiry suggests that the environment in an institution is often frustrating to individual productivity in a man's early career. There seems to be a tendency to burn up our young men and exploit their youthful energy. This may mean that, when a man reaches his forties, his capacity for further intellectual growth has been stunted.

Another vital task is the protection of research personnel during business depressions. The maintenance of key engineering personnel ought to be regarded as a "fixed charge" by any company that has research ambitions. Drastic curtailments of research and engineering staffs set a program of research back considerably more than the number of man-hours lost. Good working teams take a long time to assemble. Managements should plan for as stable a research budget as possible and not be hypersensitive to the annual profit-and-loss statement. This whole issue, of course, presents a problem which is larger than any one company or any one industry. One reason for our technological lack of preparedness in this war arose from the impact of the great depression on research and engineering. No individual company, unless it happened to be in a peculiarly favorable condition, could stand out against such a serious depression. Unfortunately, however, research departments have sometimes been the first to receive a budget cut. I hope that engineers will educate managements to an understanding of the vital importance of continuity in research programs. From the social point of view, such continuity is very much more important than continuity of dividend payments; and stockholders must, and I believe can, be educated to accept this fact.

Today, research and engineering stand high in the minds of management; and it is manpower, rather than budgetary problems, that presents the real difficulty. However, if we go through a period of difficult postwar readjustment, there is a serious danger that the tender research roots which are just beginning to sprout in the medium-sized and small companies will find themselves again deprived of nourishment.

I should like also to discuss briefly the roles of the universities and engineering schools after the war. The intensity and duration of the atomistic struggle in Europe means that the mantle of research leadership will rest on our shoulders after the war. To an extent that we frequently do not recognize, we have leaned on Europe in the past for basic and original research. Our own genius has run more to advanced engineering development. In the future the world will look to America for the most important new advances in fundamental research. The universities have a vital role to play in fundamental research, and this can be performed only with adequate financial backing. There is serious danger that they will not receive support on a scale which is adequate to our national needs and responsibilities. I would urge that the radio industry give more serious consideration to supporting university research in electronics. The comparative advantage of the universities and engineering

schools lies in fundamental research and long-range engineering development: Industry should, therefore, not be too fearful of competition from the universities, as their activities should be primarily complementary rather than competitive.

There is also need for better liaison between industrial and university research. A higher proportion of the radio companies should keep more closely in touch with the significant scientific developments that may affect their industry, particularly in the field of physics. The work of the wartime university research laboratories has made abundantly clear what can be accomplished by the application of the most advanced scientific thought to the production of new devices. After the war, could we not arrange for more exchange of scientific talent between universities and industries on short-term leaves of absence? A sabbatical year spent in an important industrial research laboratory working on a problem of keen interest to the university scientist would be an enriching experience which too few have had. Similarly for the industrial scientist, a year in a university laboratory, with some seminar teaching on the side, should be refreshing. In any event, could we not try it more?

What should the role of government be in encouraging research after the war? A radical solution would be the continuation, on a wartime scale and with government support, of the work of organizations such as the Radiation Laboratory. Such laboratories would be government institutions and might or might not be connected with a university after the war. This is the kind of plan that the Russians have adopted. It was necessary there because industry had not developed significant research of its own.

Personally, I should favor a more conservative solution. There seems no necessity for supporting government laboratories on such a scale if industry and the universities cover the field adequately. The government of the United States has a serious responsibility to maintain our technical military security; but, as I see it, the most important need for military security is to support fundamental research on a generous scale. There will be more international competition in research with Russia and with England after the war. This should be healthy, and we must not allow ourselves to fall behind. We shall have to study carefully whether some means cannot be devised, comparable in effectiveness to the operation of the Office of Scientific Research and Development during the war, whereby federal funds are made available for fundamental research in universities and nonprofit research institutes. If this is to be done, however, we must be certain that the work is carried on by first-rate minds under conditions which permit broad freedom of inquiry.

A few words also on the patent system and its relation to postwar research. I do not think that we should emasculate it. I believe that the patent system acts as an incentive to research. If there is no patent protection, industry will tend increasingly to secret practices or become primarily imitative. Nonetheless, the abuses of the patent system are very real, especially in relation to monopoly. Our antitrust legislation is not adequate to protect the public against monopolistic abuses of patents. We are in need of greater social understanding on the part of business leaders and patent holders. The patent holder of the future must be more conscious of his relations to the public. If his royalty demands are reasonable, and if his attitude is not restrictionist, there should be no serious difficulties. However, if patent holders try to perpetuate monopoly by restrictionist cross-licensing arrangements, they will play into the hands of those who want to scrap the system by legislative action. Also, if an important holder of patents refuses to pay royalties to smaller companies that are research-minded and are beginning to develop their own patent position, these companies will do everything possible to break down the patent structure of the industry. Infringement suits will abound and, with the present thinking of the courts, the patent system will soon be undermined. There is no doubt that the courts have been increasingly severe with the patent holder, but this attitude can be changed. It is monopolistic practices of which the courts have been afraid.

The electronics industry will thus need far-sighted leadership to develop the most healthy research organizations after the war. I am optimistic about this country's capacity to change its institutions to meet new situations. I believe that, twenty years from now, we shall regard our prewar organization of research and advanced engineering as very primitive. In any country that is moving forward, a dynamic and structural shift takes place in the principal occupations. Our major postwar scientific need is for more men who combine a thorough training in their subjects with creative originality. In the years immediately after the war, engineering applications of wartime developments will provide many new opportunities, and our research laboratories will be busy reducing these developments to efficient commercial practice. We must remember that we have destroyed our seed corn during the war period in that we have not been training young men for fundamental research. We are presented, therefore, with a school problem, a university problem, and an industry problem of how best to encourage and train new creative scientific talent. Industry has a very vital part to play in accomplishing this objective.

Tonal-Range and Sound-Intensity Preferences of Broadcast Listeners*

HOWARD A. CHINN†, SENIOR MEMBER, I.R.E., AND PHILIP EISENBERG†

Summary—**This paper reports the findings of an investigation to determine the tonal-range and sound-intensity preferences of a representative cross section of radio-broadcast listeners. As contrasted to former studies that have been undertaken to determine the theoretical or ideal requirements for the transmission and reproduction of broadcast program material, this undertaking ascertained the tonal range and sound intensity that the average listener considered most pleasant; that is, the method of reproduction the listener would select for use in his home when listening for enjoyment.**

Almost 500 subjects, in small groups, took part in the tests and all together, over 10,000 individual preferences were indicated. In addition to the "average" listeners, tests were undertaken with a group of professional musicians whose training presumably qualified them as critical listeners, and with a group of frequency-modulation listeners. A wide variety of program material, including popular, light-classical, and classical music, male and female vocals, and male and female spoken and dramatic speech, was presented at three tonal ranges. These were arbitrarily designated as narrow, medium, and wide. The exact frequency ranges used are shown in Fig. 1.

The trend was consistent throughout the investigation. The main conclusions of the study are:

(a) Listeners prefer either a narrow or medium tonal range to a wide one. However, the exact choice of bandwidth varies to some extent, within these limits, for different types of program content.

(b) Most listeners still prefer a narrow to a wide tonal range even when informed that one condition is "low-fidelity" and the other is "high-fidelity."

(c) Listeners prefer a peak sound intensity somewhere between 60 and 70 decibels above the acoustical reference level.

(d) Listeners prefer a higher sound intensity for speech than for music.

(e) No great differences in preferences were found between groups of different sex, age, education, and musical training; even professional musicians and frequency-modulation listeners having the same preferences.

I. Introduction

A NUMBER of investigators have undertaken studies to determine the frequency range of the human ear,[1] the manner in which this characteristic varies with age,[2] the discernibility of changes in frequency range,[3] the requirements of an ideal program-transmission system,[4] and the audible frequency ranges of musical instruments and of speech.[5] However, very little if any work has been undertaken to determine the tonal-range and sound-intensity *preferences* of a representative cross section of radio-broadcast listeners.

As contrasted to the studies that have been made, the purpose of this investigation was the determination of the method of reproduction which is most *pleasant* to the observer, rather than a study of his ability to detect changes in tonal ranges or to hear sounds of various frequencies and intensities. In addition, this study was made with a group of "average" listeners, rather than with skilled or professional observers, except, as will be noted, certain trained or experienced subjects were used in order to ascertain their reactions as contrasted to those of the "average" listener. Furthermore, the present investigation was made with a variety of both musical and voice passages under carefully controlled conditions, in order to evaluate the influence of program content upon the listeners' choice. It is believed that this study is the most extensive one of its kind that has been made to date.

II. Plan of the Study

Basic Plan:

The study consisted of two basic experiments, in each of which the participants were presented with 10 passages of speech and 10 passages of music. During each passage the form of reproduction was alternated between two different tonal ranges or two different sound intensities. For some passages, volume was kept constant while tonal range varied; for others, the tonal range was kept constant while intensity was varied, and in a few cases, as will be detailed, both factors were changed.

For each passage, the participants were asked to select from the two methods of presentation, the one they found more *pleasant* to listen to. They were asked to imagine that they were at home, listening to the radio, and to react to the broadcasts in the same way as they would at home. While the conditions of the study did not correspond exactly to home listening conditions, indications were (see Appendix I) that an overwhelming majority of the listeners would not have made a different choice if they were listening at home.

It was found necessary to conduct several types of experiments, rather than one, since preliminary studies indicated that the nature of the tests was such that fatigue or boredom might be influencing factors after the subjects had heard more than 20 to 25 passages.

Experiment I(a) was a basic experiment primarily for the purpose of determining tonal-range preferences of a cross section of listeners for male speech and classical music. Experiment I(b) involved the presentation of

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† Columbia Broadcasting System, Inc., 485 Madison Avenue, New York 22, N. Y.

[1] H. Fletcher, "Auditory patterns," *Rev. Mod. Phys.*, vol. 12, pp. 47–65; January, 1940.

[2] J. C. Steinberg, H. C. Montgomery, and M. B. Gardner, "Results of the World's Fair hearing tests," *Bell Sys. Tech. Jour.*, vol. 19, pp. 533–562; October, 1940.

[3] D. K. Gannett and I. Kerney, "Discernability of changes in program band width," *Bell Sys. Tech. Jour.*, vol. 23, pp. 1–10; January, 1944.

[4] H. Fletcher, "Hearing, the determining factor for high-fidelity transmission," Proc. I.R.E., vol. 30, pp. 266–277; June, 1942.

[5] W. B. Snow, "Audible frequency ranges of music, speech and noise," *Jour. Amer. Stand. Assoc.*, vol. 3, pp. 155–166; July, 1931.

types of program material and performers not covered by the first experiment. The program content consisted of female speech, mixed dramatic speech, piano, and popular music.

Two supplementary tonal-range experiments were undertaken with two special groups, professional musicians (Experiment II(a)) and frequency-modulation listeners (Experiment II(b)) to determine whether the basic group preferences were universal. In addition, the group of professional musicians was included in the studies to determine whether past experience with wide tonal range influenced preferences. The plan of these experiments followed exactly that of the first experiment and employed the same program material (male speech and classical music).

In the basic experiments, special electrical-transcription-type recordings were used, in order to insure uniformity and reproducibility of performance. The subjects were not advised, nor were the majority of them aware, of the fact that recordings were used. However, a supplementary tonal-range type of experiment (III) was undertaken in order to evaluate any possible influence that the use of recordings may have had upon the results. The plan of this experiment was the same as for the other tonal-range experiments except that a regular broadcast program, transmitted by wire directly from the originating studio, was used instead of recordings.

The final type of experiment undertaken (IV) was primarily for the purpose of determining sound-intensity or acoustical-level preferences. The same male-speech and classical-music selections used in the previous experiments were presented in these tests. The whole list of experiments is summarized in Table I.

TABLE I
LIST OF EXPERIMENTS

Experiment	Basic Type	Program Material	Subjects	Number of Subjects
I(a)	Tonal Range	Voice and Classical Music (recorded)	Broadcast Listener	105
I(b)	Tonal Range	Female Speech, Mixed Dramatic Speech, Piano, Popular Music (recorded)	Broadcast Listener	123
II(a)	Tonal Range	Voice and Classical Music (recorded)	Professional Musicians	20
II(b)	Tonal Range	Voice and Classical Music (recorded)	Frequency-modulation Listeners	96
III	Tonal Range	Light-Classical and Popular Music (live)	Broadcast Listener	43
IV	Sound Intensity	Voice and Classical Music (recorded)	Broadcast Listener	111
			Total	498

Subjects:

The main group of subjects, all adults, was secured by means of spot announcements broadcast over the Columbia Broadcasting System's key station, WABC, located in New York City. The professional musicians who took part in the tests were Columbia Broadcasting System staff performers. The frequency-modulation listeners were obtained from a list of frequency-modulation-set owners that had been compiled in connection with another study.

For the various experiments, a total of 29 listener sessions was conducted. In no case were the same subjects used for more than one experiment. A total of 498 listeners took part in the tests. The exact composition of the samples is detailed in Table II.

TABLE II
COMPOSITION OF THE GROUPS
(All figures are percentages)

Experiment	I(a) (Cross Section)	I(b) (Cross Section)	II(a) (Musicians)	II(b) (Freq.-Mod. Listeners)	III (Cross Section)	IV (Cross Section)
			Percentages			
Sex						
Male	48	31	100	51	30	40
Female	52	69	0	49	70	60
Age						
Under 25	49	57	0	9	38	44
Over 25	49	43	100	91	60	56
No data	2	—	—	—	2	—
Education						
Grade School	10	5	20	9	2	9
High School	67	70	55	42	67	56
College	23	25	25	47	29	34
No data	—	—	—	2	2	1
Musical Preference						
Popular	26	12	5	6	5	23
Semiclassical	47	48	0	34	42	51
Classical	27	40	95	60	51	26
No data	—	—	—	—	2	—
Musical Training						
None	28	23	0	22	33	25
Less than 1 year	30	34	0	13	7	18
1 to 2 years	24	20	0	36	20	29
More than 2 years	18	23	100*	29	40	26
No data	—	—	—	—	—	2
Instrument						
Plays none	79	76	0	53	63	66
Plays instrument	21	24	100	47	37	34
Actual Number of Persons	105	123	20	96	43	111
Number of Sessions	8	6	3	5	1	6

* 95 per cent indicated more than five years formal training.

Environment:

In order to simulate living-room conditions, at least to a degree, a small studio, having a low ceiling, was used for the study. The room used was 22 feet wide, 30 feet long, and $8\frac{1}{2}$ feet high. All properties normally associated with a broadcast studio were removed, or, as in the case of the control room, hidden from view. A rug, piano, armchairs, etc., were provided. The average noise level in the room, with no observers present, was 25 decibels above the acoustical-reference level of 10^{-16} watt per square centimeter.

The loudspeaker used to reproduce the voice and program material was located at one end of the room. It was placed behind a very light scrim curtain in order to avoid any possibility of the particular style of cabinet or the general appearance of the loudspeaker influencing the listeners.

Measurements indicated that the sound intensity in all parts of the room was essentially alike since, as detailed below, the loudspeaker had a wide angle of coverage. As an additional precaution, however, all listeners were seated towards the front of the room and as near its center line as feasible.

Equipment:

Except for one series of tests that were made with live

talent, all voice and music selections were produced from especially recorded "masters" cut on cellulose-nitrate coated disks. The records were made by the Columbia Recording Corporation, and employed the standard electrical-transcription recording characteristic.[6] This source of program material was used in order to insure absolute uniformity in the material presented at each session. Furthermore, original master recordings were used because of the extremely low surface-noise level that this type of recording affords. A new cut was used for each session to avoid any possibility of the quality being impaired by repeated playings.

The over-all response of the reproducing system was essentially uniform from 40 to 10,000 cycles per second. The background noise, even during the reproduction of the records, was negligible and not detectable by the majority of the listeners. The measured distortion of the electrical portion of the system was extremely low throughout the frequency range. Although facilities were not available to determine, quantitatively, the nonlinear distortion introduced by the loudspeaker, trained observers, critically listening to the system, agreed that no distortion of this type was detectable. This observation was consistent with the manufacturer's claims for the loudspeaker and the measured distortion of the electrical portion of the system.

The loudspeaker unit was a dual unit of well-known manufacture, employing a folded horn for the low frequencies and a multicellular horn for the high frequencies. The coverage of the latter unit was uniform over a

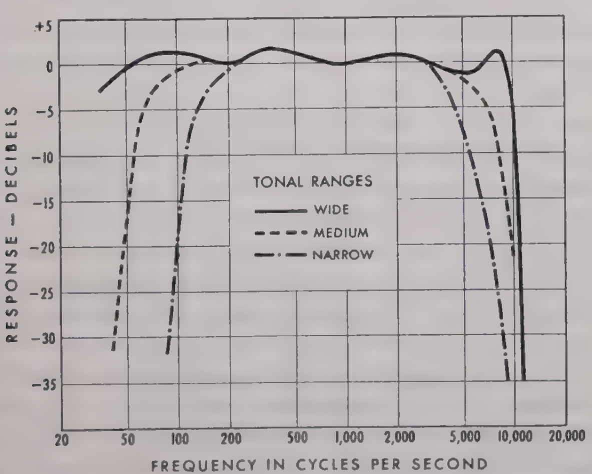

Fig. 1—The frequency limits of the three tonal ranges (wide, medium, and narrow) used for the study are shown. These response-frequency characteristics are for the entire channel, including the recording and reproducing equipment but excluding the loudspeaker.

70-degree angle in the horizontal plane and a 35-degree angle in the vertical plane. The loudspeaker was installed high enough above the floor so that listeners in the front row did not block the sound for the listeners in succeeding rows.

The system was equipped with filters to provide three tonal ranges, nominally designated narrow, medium, and wide, respectively. The filter units consisted of a single-section, prototype band-pass filter with midshunt terminations. The over-all response-frequency characteristics of the system, including the recording and reproducing equipment but exclusive of the loudspeaker, are illustrated in Fig. 1. The transmission bandwidths were chosen, in accordance with established Columbia Broadcasting System practice,[7] so as to maintain a balance between the high and the low frequencies, the product of the upper- and the lower-frequency limits, for a given attenuation, being equal to approximately 400,000. Furthermore, the band-pass filters had the characteristics shown, rather than very sharp cutoffs, since these are representative of the conditions that generally prevail in radio and recording equipment.

Program Material:

For the basic series of experiments, two classical-musical selections, Music A and B, Table III, and two speech selections, Speech A and B, were used. In order

TABLE III
List of Program Material

Selection	Title	Author	Performer	Used in Experiment
		Music Selections		
A	Carmen	Bizet	Orchestra, 40 pieces	I(a), II(a), II(b), IV
B	Preludia	Jarnefelt	Orchestra, 40 pieces	I(a), II(a), II(b), IV
C	Moon Love	McDavid	Pianist	I(b)
D	Long Ago and Far Away	Kern	Orchestra, 16 pieces	I(b)
E	An Hour Never Passes	Kennedy	Soprano and Orchestra**	III
F	If I Knew Then	Jurgens	Soprano and Orchestra**	III
G	I'll Be Seeing You	Kahal	Chorus*	III
H	Sweet and Lovely	Arnheim, Tobias, Lemas	Tenor and Orchestra**	III
I	Amor	Ruiz	Soprano and Orchestra**	III
J	Pretty Kitty Blue Eyes	Curtis	Tenor, Chorus,* Orchestra**	III
K	Long Ago and Far Away	Kern	Soprano and Orchestra**	III
L	Let Me Call You Sweetheart	Friedman	Chorus and Orchestra**	III
M	A Kiss to Remember	Silver	Soprano, Chorus,* Orchestra**	III
N	My Jesus, I Love Thee	Gordon	Soprano, Chorus,* Orchestra**	III
		Speech Selections		
A	Script "Untitled"	Corwin	Tenor Speech	I(a), II(a), II(b), IV
B	Script "Untitled"	Corwin	Tenor speech	I(a), II(a), II(b), IV
C	Script "Untitled"	Corwin	Soprano speech	I(b)
D	Mrs. Miniver	Bixby	Soprano and baritone speech	I(b)

* 14 female voices
** 29 pieces

to ascertain the effect of program content upon the results, and secondarily, the universality of the results obtained from the basic experiments, Experiment I(b) was performed with other types of program material. The supplemental selections were a piano solo covering a wide tonal range, Music C; a popular song played by a 16-piece orchestra, Music D; soprano speech, Speech C; and speech with mixed voices and sound effects, Speech D.

[6] L. C. Smeby, "Recording and reproducing standards," Proc. I.R.E., vol. 30, pp. 355–356; August, 1942.

[7] H. A. Chinn, "Broadcast studio audio-frequency systems design," Proc. I.R.E., vol. 27, pp. 83–87; February, 1939.

For the tests undertaken with live talent, Experiment III, a regular Columbia Broadcasting System broadcast program, "The American Melody Hour," was transmitted directly from the originating studio to the place where the experiment was conducted with no degradation in fidelity. The selections were of a light popular variety, played by a 29-piece orchestra with vocals by a 14-voice female chorus, a tenor, and two sopranos. The numbers are detailed in Table III, Music E to N, inclusive. During this experiment the listeners heard only the musical selections; all spoken material (announcements) was deleted.

Experimental Procedure:

The pattern followed in all listener sessions was alike. The subjects were told that, during the course of the experiment, they were to hear music and speech presented in several different ways. The listeners were advised that there were no right or wrong answers, and that they were to indicate the method of presentation they personally liked and found more pleasant. In addition, the blanks upon which the subjects recorded their choices provided means for indicating if both conditions were liked about equally well or neither was liked. Furthermore, if the listeners had a strong preference, they could so indicate.

The conditions of presentation were identified for the listeners by means of a pair of signal lights, numbered 1 and 2, which were synchronized with the changes in type of presentation. The lights were mounted vertically, one over the other.

Each test passage was exactly one minute in length, and during this period the paired conditions were alternately presented, each time for a ten-second interval. Thus, in a one-minute period, the listener had an opportunity to compare each type of presentation three times.

In the experiments, three tonal ranges and three degrees of sound intensity were used in various combinations. For convenience, these are nominally referred to herein as narrow, medium, and wide tonal ranges; and low, moderate, and high sound intensities. It is to be noted that the designations assigned are entirely abitrary. The actual tonal ranges are detailed in Fig. 1 and the sound intensity levels are given in Table IV.

TABLE IV
PEAK SOUND INTENSITIES

Nominal Designation	Sound Intensity*
Low	50 decibels
Moderate	60 decibels
High	70 decibels

* Above acoustical-reference level of 10^{-16} watt per square centimeter.

These particular limits were chosen because the "narrow" tonal range is representative of the performance of existing console type radio sets. The more popular table models have even narrower tonal ranges.[8] On the other hand, the "wide" tonal range is believed to be well above the capabilities of all radio sets with the exception of a few laboratory-type installations. The intermediate, or "medium" tonal range was chosen since, based upon liminal units,[3] it is exactly halfway between the narrow and wide tonal ranges.

In the experiments designed primarily to ascertain tonal-range preferences, five basic pairs were used for comparison as detailed in Table V. In each of the first

TABLE V
BASIC PAIRS FOR COMPARISON TESTS

Pair Number	Conditions
	Experiments I, II, and III
1	Narrow versus medium tonal range, both at moderate intensity
2	Medium versus wide tonal range, both at moderate intensity
3	Narrow versus wide tonal range, both at moderate intensity
4	Wide range at moderate intensity versus narrow range at high intensity
5	Wide range at high intensity versus narrow range at moderate intensity
	Experiment IV
6	Low versus moderate intensity, both wide tonal range
7	Moderate versus high intensity, both wide tonal range
8	Low versus moderate intensity, both narrow tonal range
9	Moderate versus high intensity, both narrow tonal range
10	Wide range at low intensity versus narrow range at moderate intensity

three comparisons, the tonal ranges are varied and the sound intensity was kept constant at a moderate, 60-decibel, acoustical level. In the last two pairs, both tonal range and sound intensity are varied.

In the sound-intensity experiment five other basic pairs were used, also detailed in Table V. In each of the first four comparisons, the volume level was varied and the tonal range kept constant; for the first two, the tonal range is wide, and for the last two, narrow. In the last comparison, both tonal range and volume are varied.

In each experiment the five comparisons were repeated for each of the four types of program content, 2 music and 2 speech, making a total of 20 pairs for each experiment, except in Experiment III, involving live talent, where only 15 pairs were presented. The order of presentation was varied and the light positions were reversed from the first music and speech selections to the second, in order to eliminate order and position effects. For instance, condition 1 for Music A became condition 2 for Music B. The purpose of repeating each comparison was to obtain an estimate of reliability, that is, the extent to which listeners made the same judgments a second time with the same material (see Appendix III).

At the completion of the main experiment, the listeners were given a brief description of low and high fidelity. They then heard two more passages, one with a wide tonal range and the other with a narrow range, with a knowledge of which signal light corresponded to each. Since the term "high fidelity" carries with it prestige that the term "low fidelity" does not, this test, in a sense, compared the influence of suggestion with the judgment of the listener's ear.

[8] O. B. Hanson, "Comments on high-fidelity," *Electronics*, vol. 17, pp. 130–131, pp. 385–391; August, 1944.

III. Analysis of Results

Tonal-Range Preferences:

There were three pairs, Table V, pairs 1 to 3, of tonal-range comparisons: (1) narrow versus medium, (2) medium versus wide, and (3) narrow versus wide. In all cases, the peak sound intensity was held constant at a 60-decibel acoustical level. Table VI presents the choices made for music and for speech for all the tonal-range experiments. Since the results for Music A and Music B are very much the same, they are combined in the table. The same is true for Speech A and Speech B.

TABLE VI
Tonal-Range Preferences

	Tonal Range	Experiment I(a) (Cross Section) Classical Music (A and B)	Experiment I(a) (Cross Section) Male Speech (A and B)	Experiment I(b) (Cross Section) Piano Music (C)	Experiment I(b) (Cross Section) Pop. Music (D)	Experiment I(b) (Cross Section) Female Speech (C)	Experiment I(b) (Cross Section) Mixed Speech (D)	Experiment II(a) (Musicians) Classical Music (A and B)	Experiment II(a) (Musicians) Male Speech (A and B)	Experiment II(b) (Frequency-Modulation Listeners) Classical Music (A and B)	Experiment II(b) (Frequency-Modulation Listeners) Male Speech (A and B)	Experiment III (Live Talent) Light-Classical Music (E to H and K to M)
						Percentages						
Pair 1	Narrow range	38	25	20	26	33	34	28	10	28	31	26
	Medium range	19	52	25	33	46	34	20	62	31	55	21
	No preference	43	23	55	41	21	32	52	28	41	14	53
Pair 2	Medium range	67	55	24	22	26	64	83	48	61	63	70
	Wide range	12	21	28	39	30	15	7	25	16	14	11
	No preference	21	24	48	39	44	21	10	27	23	23	19
Pair 3	Narrow range	58	48	30	34	29	45	73	40	59	48	40
	Wide range	15	24	24	33	34	23	5	48	28	36	26
	No preference	27	28	46	33	37	32	22	12	13	16	34
	Size of sample	105 persons		123 persons				20 persons		96 persons		43 persons

In Experiment I(a), a cross section of listeners were the subjects, and the same male-voice and classical-music passages were reproduced for all the different tonal ranges; consequently, any effect the program content had upon the preferences was the same throughout this series of tests. Here it is evident that when the narrow tonal range is compared with the medium range, pair 1, there is a preference for narrow with music and medium with speech. However, in the comparisons between medium and wide, pair 2, and between narrow and wide, pair 3, the preferences are very markedly for the narrower band for both classical music and male speech. These preferences are graphically illustrated in Fig. 2.

In Experiment I(b) a variety of program content, not covered by the preceding experiment, was presented to a cross section of listeners. In this experiment the preferences were not marked in either direction; that is, there was no distinct choice for any of the pairs in most cases. Only with mixed speech was there a marked preference for the medium over the wide range, and for the narrow over the wide range. Whereas, in the first experiments, the preference was somewhere between a narrow and a medium tonal range, in this experiment the preference seemed to fall in the medium range. It seems that the preference varies somewhat with the types of program content. In any event, in no instance is there a great preference for a wide band.

The subjects for Experiment II(a) were professional musicians, 95 per cent of whom had had more than five years' formal musical training, and all of whom had had many years of experience. The frequency-modulation listeners who were subjects of Experiment II(b) all had owned an above-average console-model frequency-modulation set for at least one year and most of them used it "frequently" or "regularly," according to their own statements.

The results for these two special groups with male speech and classical music were substantially the same as for the cross section of listeners; that is, the preference

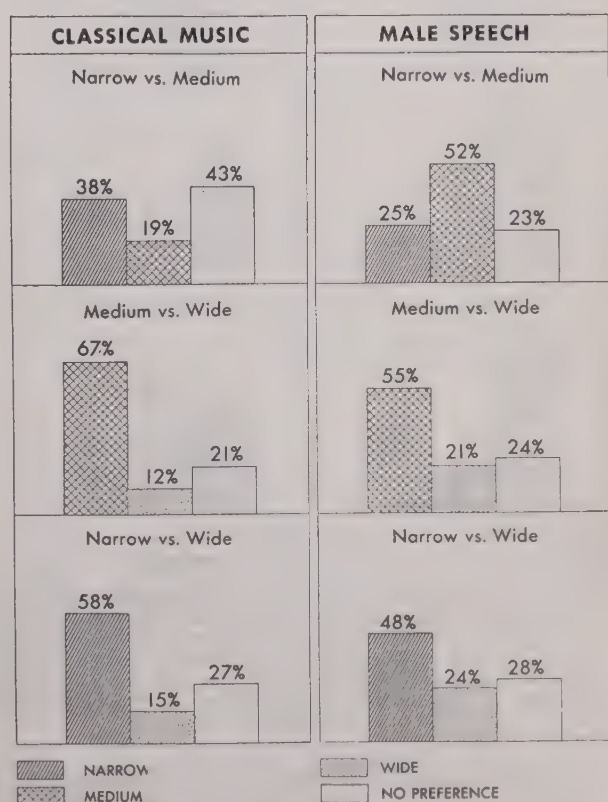

Fig. 2—The tonal-range preferences of a cross section of listeners for classical music and male speech are graphically illustrated in this chart. When the narrow tonal range was compared to the medium range, there was a preference for narrow for music and medium for speech. However, in the comparison between medium and wide and between narrow and wide, the preference was very markedly for the narrower bands for both classical music and male speech.

was for a tonal range somewhere between narrow and medium. Although there was a slight tendency for those who listen frequently to frequency modulation to prefer the medium to the narrow tonal range, the differences were not statistically significant nor were they entirely consistent. Furthermore, most of the frequency-modulation listeners, even the "regular" ones, preferred the narrow and medium bands to the wide tonal range. Thus, it seems that the preference for a narrow or medium tonal range is universal, since such choices were made even by two groups who had a greater liking for classical music than the average listener. These results for the professional musicians also tend to disprove any hypothesis that a narrow- or medium-tonal-range preference is the result of limited experience with a wide range. Fig. 3 illustrates the tonal-range preferences of the professional musicians.

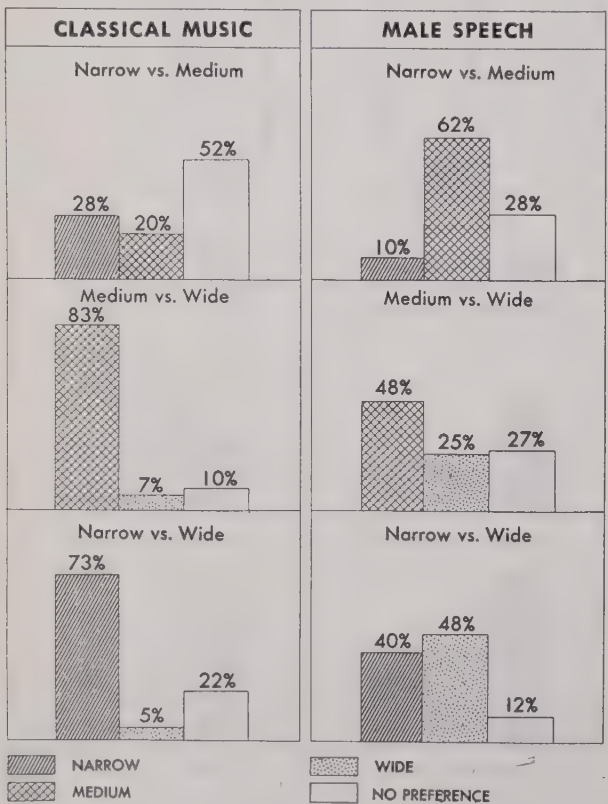

Fig. 3—The tonal-range preferences of professional musicians for classical music and male speech are illustrated in this chart. The pattern is essentially the same as for the cross section of listeners, with one exception. In the comparison between narrow and wide, the musicians had a slight preference for wide for male speech, but an even more pronounced preference for narrow for classical music than did the cross section of listeners.

In Experiment III, a cross section of listeners expressed their preferences based on tests involving live talent playing and singing light-classical and popular music. Here, when narrow and medium bands were compared, most of the judgments were equal and no marked preference for either was evident. However, as in the case of preceding experiments, there is a very marked preference for medium over wide, and a consistent preference for narrow over wide tonal ranges. It is seen, therefore, that the results obtained in the preceding experiments could not have been influenced by the use of electrical transcriptions (see Appendix IV for details).

In summarizing, the results of the above experiments indicate that listeners predominantly prefer a tonal range somewhere between a narrow and medium band. Although the preferences vary somewhat for different types of music and speech, the overwhelming majority *never* choose a wide band.

Even when the listener has been informed as to which condition is "low fidelity" (narrow tonal range) and which is "high fidelity" (wide tonal range) he still chooses low fidelity. In the playbacks of Experiments I(a), I(b) and II(a) (see Table VII) the preference for

TABLE VII
TONAL-RANGE PREFERENCES
LISTENERS INFORMED OF CONDITIONS

Tonal Range	Experiment I(a) (Cross Section)		Experiment I(b) (Cross Section)		Experiment II(a) (Musicians)		Experiment II(b) (Frequency-Modulation Listeners)	
	Classical Music (B)	Male Speech (B)	Popular Music (D)	Mixed Speech (D)	Classical Music (B)	Male Speech (B)	Classical Music (B)	Male Speech (R)
	Percentages							
Narrow	55	55	34	44	80	63	31	48
Wide	32	34	37	34	15	32	55	46
No Preference	13	11	29	22	5	5	14	6
Size of sample	105 persons		123 persons		20 musicians		96 persons	

the narrow band is as great when the listeners were informed of the conditions as when they were not. However, there is a slightly greater preference for the wide band, but this is at the expense of the equal judgments. In all probability, the increased preference for the wide band indicates that some of the subjects were impressed by the term "high fidelity."

The frequency-modulation listeners showed a greater preference for "high fidelity" over the "low" than any other group when informed of the conditions (Experiment II(b)). In fact, the more they listen to frequency-modulation, according to their own statements, the more they indicated preference for high fidelity *after* they had been acquainted with the conditions of the tests. Since these special listeners did not show these preferences when they were unaware of the conditions of the tests, it seems that they were more influenced by the prestige of the term "high fidelity" than any other group.

A point that lends further support to this conclusion is that frequency-modulation listeners, after being informed of the conditions, show an even more marked preference for the "high-fidelity" condition for music than for speech. These preferences are the reverse of those made *before* the subjects were informed of the conditions. This seems to indicate that frequency-modulation-set owners have been susceptible to the various forms of publicity that stress high quality for musical programs on frequency-modulation. The "uninformed"

and "informed" preferences of frequency-modulation listeners as compared with a cross section of listeners are illustrated in Fig. 4.

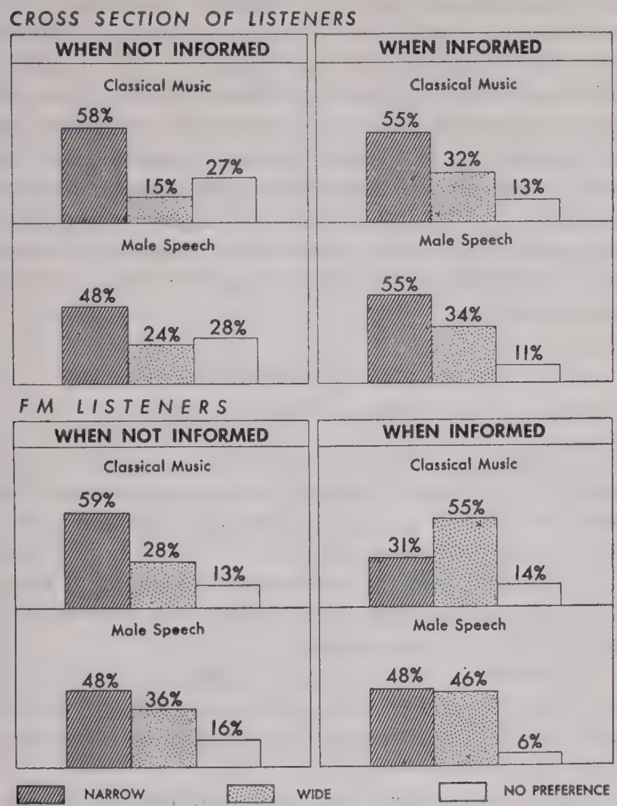

Fig. 4—When informed of the conditions of the test, the cross section of listeners had a slightly greater preference for wide range than when not informed. However, the shift was almost entirely at the expense of the group that indicated "no preference" when the conditions were unknown to them. For male speech, the frequency-modulation listeners' preferences changed in a similar manner when informed of the conditions. On the other hand, for classical music, the frequency-modulation listeners reversed their preference when informed of the conditions of this test. This was the only group where the prestige of the term "wide range" or "high fidelity" appeared to counteract the original unbiased choice of the listener.

Sound-Intensity Preferences:

In Experiment IV, there were four pairs, Table V, pairs 6 to 9, of volume-level comparisons; between low and moderate, and between moderate and high sound intensities, first at a wide tonal range and then at a narrow tonal range. As before, the results for both Music A and B and for both Speech A and B were combined. These results are presented in Table VIII.

TABLE VIII
SOUND-INTENSITY PREFERENCES

	Sound Intensity	Tonal Range	Experiment IV Classical Music (A and B)	Experiment IV Male Speech (A and B)
			Percentages	
Pair 6	Low	Wide	9	2
	Moderate	Wide	72	79
	No Preference		19	19
Pair 7	Moderate	Wide	27	42
	High	Wide	46	38
	No Preference		27	20
Pair 8	Low	Narrow	18	4
	Moderate	Narrow	66	91
	No Preference		16	5
Pair 9	Moderate	Narrow	49	28
	High	Narrow	30	52
	No Preference		21	20

(Size of sample: 111 persons)

There is a marked preference for the moderate, 60-decibel sound-intensity level as against the low, 50-decibel level either at a narrow or a wide tonal range, pairs 6 and 8. But there is no clear choice between the moderate and the high, 70-decibel levels, pairs 7 and 9. In other words, the sound-intensity preference for most listeners seems to lie somewhere between 60 and 70 decibels.

In general, the sound-intensity preference for speech seems to be somewhat louder than for music, particularly for narrow-band reproduction, pairs 8 and 9.

Tonal-Range and Sound-Intensity Preferences:

To complete the study, experiments were undertaken to learn what the preferences would be when both tonal

TABLE IX
TONAL RANGE AND SOUND-INTENSITY PREFERENCES

	Tonal Range	Sound Intensity	Experiment I(a) (Cross Section) Classical Music (A and B)	Experiment I(a) (Cross Section) Male Speech (A and B)	Experiment I(b) (Cross Section) Piano Music (C)	Experiment I(b) (Cross Section) Pop. Music (D)	Experiment I(b) (Cross Section) Female Speech (C)	Experiment I(b) (Cross Section) Mixed Speech (D)	Experiment II(a) (Musicians) Classical Music (A and B)	Experiment II(a) (Musicians) Male Speech (A and B)	Experiment II(b) (Frequency-Modulation Listeners) Classical Music (A and B)	Experiment II(b) (Frequency-Modulation Listeners) Male Speech (A and B)	Experiment III (Live Talent) Light-Classcial Music (I to K, M and N)
			Percentages										
Pair 4	Wide	Moderate	21	18	42	41	35	24	3	5	13	19	22
	Narrow	High	62	70	51	38	52	61	67	75	79	72	73
	No Preference		17	12	7	21	13	15	30	20	8	9	5
Pair 5	Wide	High	28	43	55	18	41	46	10	30	40	45	47
	Narrow	Moderate	56	34	32	67	37	39	38	28	48	39	37
	No Preference		16	43	13	15	22	15	52	42	12	16	16
	Size of Sample		105 persons		123 persons				20 persons		96 persons		43 persons

	Tonal Range	Sound Intensity	Experiment IV (Cross Section) Classical Music (A and B)	Experiment IV (Cross Section) Male Speech (A and B)
			Percentages	
Pair 10	Wide	Low	10	2
	Narrow	Moderate	76	90
	No Preference		14	8
	Size of Sample		111 persons	

range and sound intensity were varied simultaneously. As a part of Experiments I(a), I(b), II(a), II(b) and III, two such comparisons were undertaken, Table V, pairs 4 and 5; and in Experiment IV, one comparison, Table V, pair 10. The results of this study are given in Table IX.

It is evident that, in all three comparisons, the narrow tonal range is preferred to the wide. However, the strength of the preferences depends upon the mixture of the elements. When (pair 10) narrow tonal range at moderate volume, both of which are preferred elements, is compared with wide tonal range at low intensity, both of which are unpreferred elements, then the choice is overwhelmingly for the preferred combination. In the other cases, where the comparisons were not between definitely preferred and unpreferred elements, the choices were not as consistent or as clear-cut.

TABLE X
TYPICAL GROUP PREFERENCES
(Combined Data for Classical Music and Male Speech)

Groups	Tonal-Range Experiment I(a)			Sound-Intensity Experiment IV		
	Narrow Tonal Range 60-decibel level	Wide Tonal Range 60-decibel level	No Preference	Narrow Tonal Range 60-decibel level	Narrow Tonal Range 70-decible level	No Preference
	Percentages					
Sex						
Male	57	17	26	41	42	17
Female	49	22	29	35	42	23
Age						
Under 26	55	19	26	34	41	25
26 and over	52	20	28	42	41	17
Education						
Grade School	46	18	36	48	37	15
High School	51	21	28	32	47	21
College	60	17	23	46	33	21
Musical Training						
None	40	25	35	39	40	21
Less than 1 year	56	19	25	26	55	19
1 to 2 years	56	14	30	39	36	25
2 or more years	61	18	21	43	41	16
Musical Preference						
Popular	55	20	25	40	42	18
Semiclassical	49	19	32	39	42	19
Classical	57	21	22	35	40	25
Instrument						
Plays none	52	18	30	38	42	20
Plays	56	24	20	39	39	22

Group Differences in Preference:

The preferences of the listeners of Experiment I(a) and IV were analyzed by sex, age, education, amount of musical training, musical preferences, and whether or not they played an instrument. The criterion of a difference was determined not only by the size of the difference but also by its consistency.

In general, the analyses reveal no differences for any of these breakdowns, with one possible exception; women display a consistent but *very slight* tendency to prefer wider tonal range than men. This trend is somewhat compensated for by a greater amount of "no preferences" by women.

Table X shows the results for a typical tonal-range passage from Experiment I(a) and a typical volume level passage from Experiment IV. In this table all types of content are combined (Music A and B and Speech A and B) because the results for each are very similar.

It is interesting that those subjects over 26 years of age, when the average listener begins to suffer a hearing loss,[2] show no difference in preference from younger listeners. Education plays no apparent role in preference, nor does musical experience. Furthermore, the results of the tests with the professional musicians and the frequency-modulation listeners substantiate the finding that group differences, such as have been investigated, do not influence tonal-range or sound-intensity preferences. Such uniformity in results for various types of groups indicates no need for larger samples than were used.

IV. Conclusions

The main conclusions of this study are:

1. Listeners prefer either a narrow or medium tonal range to a wide one. However, the exact choice of band width varies to some extent within these limits, for different types of program content.

2. Listeners prefer a narrow to a wide tonal range, even when informed that one condition is "low fidelity" and the other is "high fidelity."

3. Listeners prefer a slightly wider band for female speech, piano, and popular orchestra selections than for male speech, mixed dramatic speech, and classical orchestra selections.

4. Listeners prefer a peak sound-intensity level somewhere between 60 and 70 decibels above the acoustical-reference level.

5. Listeners prefer higher sound intensity for male speech than for classical music.

6. Both tonal range and sound intensity influence the listeners' choices. The most preferred combination, within the limits of this study, seems to be a narrow tonal range and a 60- to 70-decibel acoustical level.

7. The reasons given for the choices made indicate that the listeners were not making analytical judgments, but rather, were judging in terms of what was pleasant to their ears. In short, they were judging naturally, as they would at home (see Appendix I).

8. There were no clear-cut and consistent differences between various subgroups of the audience; sex, age, education, amount of musical training, musical preferences, or whether they played a musical instrument.

9. Two groups with special experience, professional musicians and frequency-modulation listeners, indicated substantially the same preferences as did the cross section of listeners.

V. Discussion

The main conclusion derived from these experiments was that listeners prefer either a narrow or medium tonal-range band to a wide band when listening to reproduced sound. It may be argued, however, that this

preference for restricted tonal range is the result of years of experience of listening to a narrow band on the radio and on recordings. The argument generally continues that, with experience, listeners will overcome this initial dislike and discover new enjoyment in wide range. It is important to understand that this argument is based on no published experimental evidence whatsoever. It is an hypothesis unfortunately very often stated as a fact. Actually, listeners have had extensive experience with wide range in speech, because that is what they hear all day long in ordinary conversation, yet in these tests, they did *not* prefer a wide band for speech. In addition, listeners have experienced in the sound produced in motion-picture theaters a relatively wide range.

With regard to music, it has been found that listeners prefer narrow range, regardless of how much musical training and experience they have had. Even professional musicians, who have had long years of experience with music produced at the widest natural range, still manifest the same preferences as other listeners.

It is admitted that these experiments do not furnish conclusive evidence to settle this question one way or the other. The data, however, tend to confirm the alternative hypothesis; that a narrower tonal range is preferred because it sounds better, and not because the listeners' tastes have been spoiled.

The factor of suggestion is also an important consideration in this connection. It was found in these experiments that, although listeners do not prefer a wide tonal range, the entire matter of range is probably not as important to them as is sound-intensity level. When intensity level was controlled, the choices were not made with great confidence as shown by the relatively small amount of strong preferences, by the high percentage of no preference judgments, and by the frequent changes of judgment on repeat tests. Apparently, then, listeners cannot discriminate as well between differences in tonal range as they can between sound-intensity levels. Certainly tonal range is not as important as freedom from background noise. Because tonal range is generally not a matter of great concern to most listeners, and because strong preferences are not evident, it was possible to influence some listeners to choose a wide range when they were told the presentation was "high fidelity." On the other hand, it is unlikely that verbal suggestions would influence listeners to choose sound-intensity levels they disliked or presentations with excessive background noise.

If, at some future time, there should be a predominant shift to a preference for wide range, the question may well be raised as to the cause of the change. Would it be the result of greater listening pleasure through wide-range experience, or would it be simply the prestige-suggestion effect resulting from the use of the terms "wide range" and "high fidelity"?

* * *

This paper presents, without bias, the results of a scientific investigation. Consequently, the conclusions reached should not be construed as an indication of the possible trend of any official policies the Columbia Broadcasting System may adopt on matters upon which the subject may have a bearing. Furthermore, it is recognized that this study covers only a few of the many ramifications of the general problem. Further work may be undertaken and reported upon.

Appendix I

Influence of Test Environment

A technical question of considerable interest in connection with the results that have been presented is the extent to which these results would have been different if it had been possible to duplicate the home situation. To answer this problem, each listener was asked the question: "Do you think that you would have made different choices if you had been listening at home?" The answers were as tabulated in Table XI.

TABLE XI

Differences from Home Listening

"Do you think that you would have made different choices if you had been listening at home?"

	Experiment I(a) (Cross Section)	Experiment I(b) (Cross Section)	Experiment II(a) (Musicians)	Experiment II(b) (Frequency-Modulation Listeners)	Experiment III (Live Talent)	Experiment IV (Cross Section)
	Percentages					
Yes	11	14	20	9	12	5
No	80	67	60	73	77	82
Don't know	9	19	20	18	11	13

It is seen that in all cases an overwhelming majority indicated that the conditions of the experiment did not violate the home-listening experience in any crucial way.

In addition, the listeners were asked to indicate the reasons for their choices during the experiment. Most of the reasons given were that one condition was pleasanter than the other, or clearer, or more natural, or richer. That these terms are relatively interchangeable is indicated in Experiment I(a) by the fact that only 39 per cent of the reasons given were identical from one experiment to another. When this figure is compared with the reliability of identical choices, Table XIII, which averages well above this figure, it is seen that the reliability of "reasons" is low.

When the reasons given are broken down by actual preferences, relatively the same distributions are found. This indicates only that the listeners were making a natural and unanalytical judgment; that is, they were reacting as they would at home. They judged by what sounded pleasantest to their ears.

Appendix II

Effect of Signal-Light Positions

A second technical question of interest, in view of results obtained by other investigators, was the extent to

which the position of the signal lights themselves may have influenced the judgments made.

As already detailed, the signal lights were mounted vertically, one over the other, the upper light being No. 1, the lower, No. 2. This was done to obtain data that might afford an interesting contrast to a former study[3] wherein there was a marked preference for the right-hand signal position when there was little or no difference in the two samples being presented.

TABLE XII

EFFECT OF SIGNAL-LIGHT POSITIONS

Passages	Light 1	Light 2	Equal
	Experiment I(a)		
		Percentages	
Music A and Speech A	26	51	23
Music B and Speech B	47	29	24
	Experiment I(b)		
Music C and Speech C	33	36	31
Music D and Speech D	39	35	26
	Experiment II(a)		
Music A and Speech A	27	50	23
Music B and Speech B	43	22	35
	Experiment II(b)		
Music A and Speech A	31	52	15
Music B and Speech B	52	28	20
	Experiment III		
Number 1	27	54	19
Number 2	55	20	25
Number 3	22	45	33
	Experiment IV		
Music A and Speech A	17	69	14
Music B and Speech B	59	21	20

In each experiment, the signal lights were reversed for the second music-and-speech passage, as compared to the first. That is, what was condition 1 for Music A and Speech A, became condition 2 for Music B and Speech B; similarly, for Music C and Speech C versus Music D and Speech D. In Experiment III, the lights were reversed in passage No. 2 as compared to passage Nos. 1 and 3.

Table XII is an analysis of the results, and indicates very clearly that the choices were made in terms of what the listener heard, rather than because of any preferences for a light number or position. In every case, when the lights were reversed, the judgments followed accordingly.

APPENDIX III

CONFIDENCE AND RELIABILITY OF JUDGMENTS

A third technical question is the confidence in and reliability of judgments. Three figures were found to be useful indexes: (1) extent of strong preferences, (2) extent of equal judgments (no preferences), and (3) degree of identity of choices for like conditions with the different sets of music and voice.

It was found that strong preferences and identity of choice were positively related to each other; that is, the greater the strong preferences, the greater the identity of choice. The extent of equal judgments was negatively related to the other two indexes; that is, the greater the equal judgments, the fewer the strong preferences, and the smaller the identity of choice.

Table XIII presents the results for music and speech by tonal comparisons, sound-intensity comparisons, and by the combined type of comparisons.

The tests wherein both the sound intensity and the tonal range were varied were the most reliable. The subjects indicated strong preferences more often, indicated no preference less often, and made the same judgment on repeated tests more often. This indicates in another way that both sound intensity and tonal range influence the judgment. When both factors vary, it is easier to make a reliable judgment than when only one factor varies. However, the sound-intensity judgments were more reliable than the tonal-range judgments. This seems to indicate that, although both factors influence the judgment, listeners can distinguish more readily between different sound intensities than between different tonal ranges.

The judgments for Experiment I(b), Variety of Content, and Experiment III, Broadcast Music, were less reliable than in the other experiments. That is, there were fewer strong preferences, more equal judgments, and less identity of choice. In the broadcast-music experiment this could be accounted for because of lack of control of content. However, in the variety-of-content experiment, although different types of material were presented, it was controlled in the same way as in Experiments I(a) and II(a).

TABLE XIII

INDEXES OF CONFIDENCE AND RELIABILITY

Test	Experiment I(a) (Cross Section)		(Experiment I(b) (Cross Section)		Experiment II(a) (Musicians)		Experiment II(b) (Frequency-Modulation Listeners)		Experiment III (Live)	Experiment IV (Cross Section)	
	Music	Speech	Music	Speech	Music	Speech	Music	Speech	Music	Music	Speech
					Percentage of Strong Preferences						
Sound Intensity	—	—	—	—	—	—	—	—	—	12	22
Tonal Range	13	12	5	9	10	8	9	16	7	—	—
Intensity-Tonal Range	22	17	20	11	14	19	32	27	17	19	27
					Percentages of No Preferences						
Sound Intensity	—	—	—	—	—	—	—	—	—	21	16
Tonal Range	30	25	44	31	28	23	26	18	35	—	—
Intensity-Tonal Range	16	18	14	16	41	31	10	13	11	14	8
					Percentage of Identical Choices						
Sound Intensity	—	—	—	—	—	—	—	—	—	48	63
Tonal Range	58	55	40	40	52	47	51	56	44	—	—
Intensity-Tonal Range	59	64	47	63	63	63	70	71	58	65	80

A possible explanation is that the types of music and speech used influenced the judgments. In Experiments I(a) and II(a), a slightly wider tonal range was preferred for male speech than for classical music. Apparently, an even wider tonal range was preferred for the two types of music and the female speech used in the variety-of-content experiment, although no preference was shown for the widest range.

It seems that, as the tonal-range preference approached medium, the judgments became less reliable since the listener did not have clear preferences. Further evidence for this view is obtained from the fact that judgments of male speech, which was closer to a medium band in preference than classical music, were slightly less reliable than judgments of classical music.

Appendix IV

Live Talent Versus Recorded Program Material

Except for the fact that the reliability of judgments was somewhat less (see Table XIII) for Experiment III, which employed live talent instead of recorded program material, the tonal-range preferences were consistent with all other experiments (see Table VI). The lower reliability seemed to be due to the conditions of the experiment rather than any basic difference between judgments of recordings as compared with live talent. In the live-talent experiment the program content varied and, as already discussed in Appendix III, this influenced the listeners to some extent.

However, since the same results were obtained with both live talent and transcription, it follows that experiments of this kind can safely be conducted with low-surface-noise, electrical-transcription-type recordings. In fact, the use of recordings is preferred since the control of content made possible with records increases the reliability of the judgments.

Exalted-Carrier Amplitude- and Phase-Modulation Reception*

MURRAY G. CROSBY†, FELLOW, I.R.E.

Summary—**An amplitude- or phase-modulation receiving system is described in which the harmonic distortion produced by fading of the carrier with respect to the sidebands is eliminated. The various parts of such a receiver, including the carrier filter, automatic-frequency-control discriminator, and detecting systems, are described. Analyses are given of the selectivity effect due to carrier exaltation and of exalted-carrier diode and multigrid detection. The optimum degree of carrier exaltation and the effect of carrier limiting are discussed. Results are given of observations of reception on an exalted-carrier diversity receiving system.**

Introduction

THE RECEPTION of amplitude- or phase-modulated signals which have been transmitted via the ionosphere is usually marred by multipath transmission effects which produce selective fading. The most destructive effect of the selective fading is the fading of the carrier frequency with respect to the sideband frequencies. This produces an overmodulation which results in harmonic distortion usually consisting of a predominance of second harmonic, but there may also be intermodulation components of the modulating wave included. Such fading is common in medium-frequency broadcasting at the region where the sky wave and ground wave are approximately of equal strength. This condition occurs at night in the region from 50 to 150 miles from the transmitter. In high-frequency transmission, where the ionosphere is depended upon for transmission, carrier fading is practically always present.

It is the purpose of this paper to describe systems for receiving amplitude- and phase-modulated waves in which the harmonic distortion due to carrier fading is eliminated. The elimination of the distortion is brought about by filtering the carrier and recombining it with the signal at a raised or exalted level, or by recombining in a type of detector which inherently eliminates the carrier-fading distortion.

The use of a filtered and reintroduced carrier has been the practice in the reception of single-sideband telephony for some time.[1] In this prior use, the primary function of raising the carrier at the receiver is that of bringing the carrier back up to its proper strength from the reduced value radiated by the transmitter. By intentionally reducing the carrier at the transmitter, a power gain is effected since the power capabilities of the transmitter may be concentrated on the sidebands.

In the systems of this paper, the conventional double-sideband amplitude modulation with normal carrier radiation is assumed. Since the receivers may be adapted to the reception of phase modulation by a simple change in the phase of carrier recombination, phase modulation is also considered. When phase modulation is received in this manner, a maximum phase deviation of approximately one radian may be accommodated. This degree of phase deviation produces sidebands which occupy about the same frequency spectrum as double-sideband amplitude modulation. The reception of this type of

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† Paul Godley Company, Upper Montclair, N. J. The work on this paper was done at R.C.A. Laboratories, Radio Corporation of America, Riverhead, L. I., N. Y.

[1] F. A. Polkinghorn and N. F. Schlaack, "A single-side-band short-wave system for transatlantic telephony," Proc. I.R.E., vol. 23, pp. 701–718; July, 1935.

phase modulation has been considered previously.[2] The present paper describes improved methods of reception which are readily adapted to either phase- or amplitude-modulation reception by the exalted-carrier method.

The Receiver Circuit

The block diagram of Fig. 1 shows the essential elements of an exalted-carrier receiver. Units 1 to 7, inclusive, may comprise a conventional double-superheterodyne receiver. The intermediate-frequency output of this receiver is divided into two branches. One of these branches feeds the recombining detector 12 directly. The other branch feeds carrier filter 9, which separates the carrier from the sidebands. The filtered carrier is fed from 9 to limiter 10, which maintains the carrier amplitude at a constant value. The filtered and limited carrier is then recombined with the unfiltered signal at a phase determined by phase adjuster 11. The resulting combined signal is detected by detectors in 12 and fed to audio-frequency amplifier 13 for utilization.

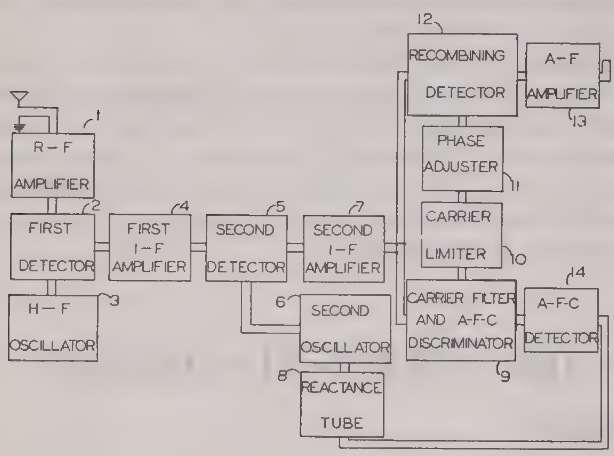

Fig. 1—Block diagram of an exalted-carrier receiving system.

The most important part of the carrier-exalting circuit is the carrier filter and automatic-frequency-control discriminator which is contained in unit 9 of Fig. 1. In previous circuits of this type,[2] an off-neutralized crystal filter of the three-electrode type was arranged in a circuit to provide automatic-frequency-control potentials and phase-modulation detection from the same crystal filter and detectors. Amplitude-modulation detection was accomplished on a separate pair of detectors. The circuits described here are somewhat similar, but the crystal-filter circuit is arranged to provide two outputs; namely, an output of pure carrier without sidebands, and a pair of outputs which may be fed to differential detectors to generate automatic-frequency-control potentials. The pure carrier output is fed to a separate detector where it is recombined with the signal for exalted-carrier detection.

An important feature of the carrier-filter system is the provision of an automatic-frequency-control discriminator which utilizes the same crystal filter for both the carrier filter and the frequency discriminator. This makes unnecessary the careful synchronization that would be required if the two functions employed separate crystal filters or other separate automatic-frequency-control circuits. In the circuits to be described here, the crystal-filter output is arranged so that pure carrier is obtained from one output terminal and automatic-frequency-control discrimination from other terminals. This is done with a single two-electrode type of quartz crystal.

The circuit of Fig. 1 shows the use of a double-intermediate-frequency superheterodyne system for the convenience that is afforded in the application of automatic frequency control. Application of the control to the second oscillator which is fixed in frequency simplifies the automatic-frequency-control system in that the degree of control is fixed. A single intermediate-frequency system could likewise be used by applying the automatic-frequency control to a reactance tube on the first oscillator 3, and feeding the carrier-exalting and detecting circuits directly from the first and only intermediate frequency.

Carrier Filter and Automatic-Frequency-Control Discriminator

Fig. 2 shows a form of the carrier filter and automatic-frequency-control discriminator circuit.[3] A crystal filter of the type commonly used in communications-type receivers serves as the carrier filter. The output

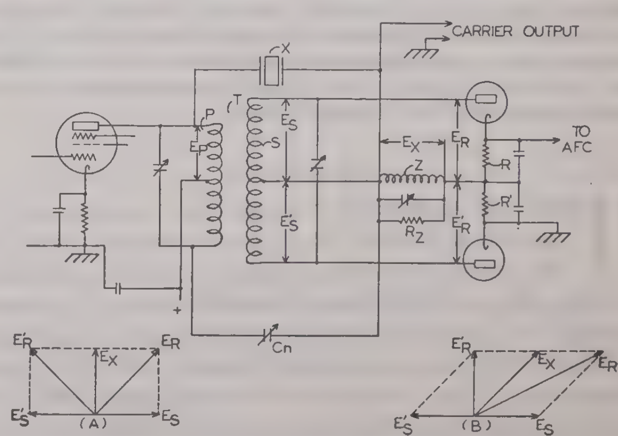

Fig. 2—Carrier filter and automatic-frequency-control discriminator circuit.

of the carrier filter is also combined with the unfiltered energy in proper phase so as to form a sharp frequency discriminator similar to the phase-shift type of discriminator used in frequency-modulation reception. The crystal X is fed by midtapped primary P, which also furnishes opposite-phase voltage to capacitor C_n for neutralizing the electrode capacitance of the crystal holder. The output of the carrier filter appears across

[2] Murray G. Crosby, "Communication by phase modulation," Proc. I.R.E., vol. 27, pp. 126–136; February, 1939.

[3] Murray G. Crosby, New Zealand Patent No. 88,212, "Phase-modulation detectors," issued March 27, 1944.

tuned circuit Z, which is normally tuned to the crystal frequency. This output is tapped at this point to provide pure carrier for recombination with the signal.

The filtered output also combines with the secondary voltages from transformer T. Diagram (A) in Fig. 2 shows vectorially how the crystal output combines with the secondary voltages from transformer T. The primary voltage E_p is not shifted in phase by the crystal filter since the crystal is operated at series resonance and Z is tuned to resonance. Hence the phase of E_x is the same as that of E_p. The secondary voltages E_s and E_s' are shifted 90 degrees, with respect to the primary voltage, by the phase shift inherent in tuned transformer T. For the in-tune condition shown in Diagram (A), the resultant voltages E_R and E_R' are equal in amplitude at the detector inputs. The voltage rectified by the diodes and appearing on the differentially-connected diode resistors R and R' will therefore balance so that the automatic-frequency-control output potential is zero. When the carrier frequency is out of tune with the crystal, the phase of the crystal output is shifted so that vector diagram (B) is produced. For this condition, the resultant voltages fed to the differential detectors are out of balance so that a difference voltage appears across the diode resistors. This voltage is positive or negative depending upon the direction of the frequency shift. The result is a frequency-discriminator characteristic as shown in Fig. 3 which has its midpoint at the filtered carrier frequency F_c and has a sensitivity in accordance with the selectivity of the crystal filter. Fig. 3 is an actual characteristic taken on an experimental exalted-carrier receiver.

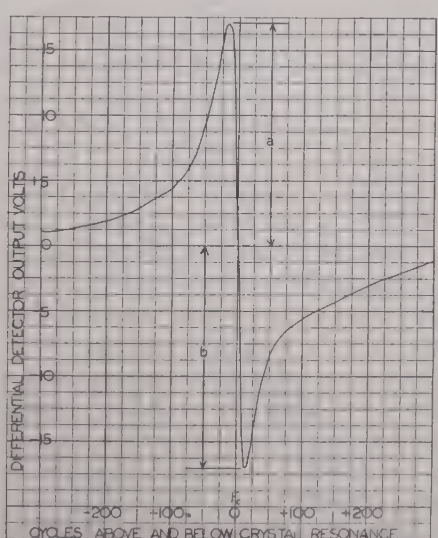

Fig. 3—Automatic-frequency-control detector output characteristic.

The degree of selectivity of the carrier filter and sharpness of the automatic-frequency-control discriminator may be controlled by a variation of resistor R_Z. The selectivity increases as R_Z is made lower, in order to lower the output impedance of the crystal filter. A variable resistance in series with the inductance or capacitance of tuned circuit Z may also be used. This variation of the crystal-output impedance controls the selectivity in a manner well known in the use of this type of crystal filter in communications-type receivers.[4]

Various other circuit modifications are possible. The essential requirement is the combination of the filtered and unfiltered signals at a phase difference of 90 degrees and in a manner such that the output of the carrier filter may be tapped at a point where the unfiltered signal does not appear. Fig. 4 shows a circuit arrangement in which all of the driving voltage is obtained from a midtapped secondary. The crystal-filter output is neutralized in the usual manner and is combined with the unfiltered signal by means of isolating and phase-shifting impedances C_3, C_3', C_4, and C_1, R_1, C_2, R_2.

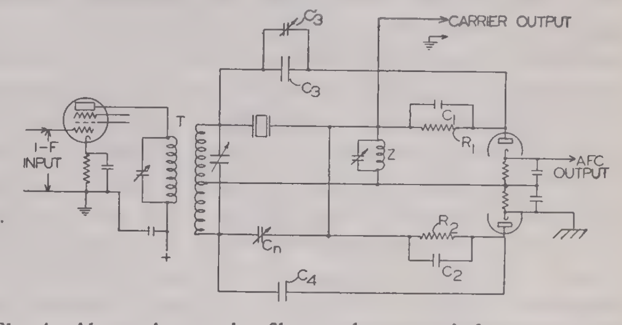

Fig. 4—Alternative carrier filter and automatic-frequency-control discriminator circuit.

Assuming, for the moment, that capacitors C_1 and C_2 are omitted, it can be seen that, due to the resistive crystal-output circuit, the phase of filtered voltage fed to the detector is unchanged at carrier frequency. The unfiltered signal, which is fed to the detectors through C_3, C_3', and C_4, is shifted somewhat less than 90 degrees since it is a capacitance feed to a resistive detector input. C_3' is a trimmer for equalizing the magnitude of $C_3'+C_3'$ and C_4. Final adjustment of the phase is made by slightly detuning tuned circuit Z. This detuning increases the effective selectivity, or Q, of the crystal filter by providing a lower impedance in the same manner as the control of Q effected by the variation of resistor R_Z in Fig. 2. Still higher values of Q may be effected by adding capacitor C_1 and C_2 to the filter-output coupling circuits. These capacitors shift the relative phase of the filtered and unfiltered signals in a manner to require a further detuning of Z to provide the proper 90-degree phase relation. The result is a means of controlling the selectivity of the crystal filter. The selectivity or Q is minimum when C_1 and C_2 are omitted and Z is tuned nearest to resonance.

In aligning the crystal discriminator, the condition of the proper phase relation between the filtered and unfiltered voltages is conveniently indicated by an equality between the maximum positive and negative swings from the differentially detected output of the discriminator (equality of a and b in Fig. 3). Tuned circuit Z is

[4] D. K. Oram, "Full-range selectivity with 455-kc quartz crystal filters," *QST*, p. 33; December, 1938.

therefore tuned to effect this equality. Balancing capacitor C_3' is adjusted so that the automatic-frequency-control potentials do not reverse polarity at frequencies other than the carrier frequency.

The circuit of Fig. 4 has the added advantage that the second harmonic of the rectified-signal component from the push-pull secondary of transformer T does not appear across the filter-output impedance Z. In the circuit of Fig. 2, impedance Z is in the common leg of what amounts to a full-wave rectifier circuit which places full-wave rectified half waves of the unfiltered signal across impedance Z. In spite of the fact that Z is tuned to the crystal frequency, enough second harmonic of the signal frequency may appear across Z to upset measurements of the selectivity of the crystal filter. Subsequent filtering may be used to remove the second harmonic when the circuit of Fig. 2 is used or the harmonic may be eliminated by reversing one of the diodes so that the rectification of the signal from the push-pull transformer is half-wave instead of full-wave.[3] This latter alternative requires rearranging the diode resistors to effect differential detection.

The selectivity of the carrier filter may either be adjusted to remove the sidebands completely so that limiting is not required for this function, or merely enough selectivity may be used to keep amplitude modulation at the carrier-limiter input somewhat below 100 per cent. In the latter case, the limiter is depended upon to remove the remaining amplitude modulation. A crystal filter with an equivalent Q of 2000 at an intermediate frequency of 50 kilocycles has been found sufficient to remove the sidebands of program modulaton. In amplitude-modulation reception, selectivities less than this require the use of a carrier limiter to prevent overemphasis of the low-modulation frequencies which pass the carrier filter. In phase-modulation reception, the carrier limiter cannot remove modulation sidebands so that lower selectivities result in a reduced output from the low-modulation frequencies which are passed by the carrier filter. This is brought about by the fact that, unless the sidebands of modulation are removed from the carrier channel, there is no relative phase shift between the filtered-carrier channel and the unfiltered-signal channel to effect detection.

Single-Diode Detection

The primary objective to be accomplished by the recombination and detection of the filtered carrier and unfiltered signal is that of eliminating the distortion that is brought about by the condition of overmodulation that results from carrier fading. One method of effecting this result is by feeding the carrier to a single-diode detector with the amplitude of the carrier exalted with respect to the unfiltered signal. This reduces the effective percentage of modulation fed to the detector so that the carrier may fade to a greater depth before overmodulation occurs. In other types of detectors, the distortion products are either balanced out or are not generated. Three types of these kinds of detection will be considered here: (1) single-diode detection; (2) balanced-diode detection; (3) multigrid detection. The effect of limiting the carrier will also be considered.

Fig. 5 shows a circuit for combining the filtered carrier and unfiltered signal for detection on a single diode. Tube A may function as either a carrier amplifier or limiter, but it will be assumed to be an amplifier for this discussion. Transformer T, which feeds diode C, is common to the output of tube A and the unfiltered signal amplifier B. The phase of combination is adjusted by shifting the phase of the unfiltered signal by means

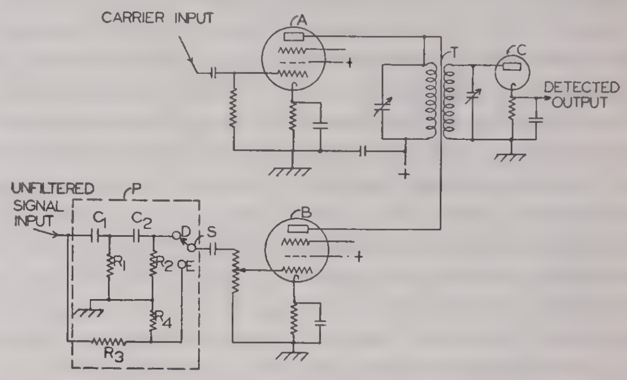

Fig. 5—Single-diode detector circuit.

of phase shifter P. It is sometimes more feasible to place the phase shifter in the unfiltered-signal branch as shown. This removes the attenuation effected by the phase shifter from the carrier branch which is usually fed to the detector with as high a level as possible. Fixed values of phase shift are chosen by means of switch S. For instance, point D on switch S might give the proper phase for phase-modulation detection, and point E for amplitude-modulation detection. Resistors R_3 and R_4 effect an attenuation equal to that produced by phase-shifting elements C_1, R_1, and C_2, R_2.

For amplitude-modulation detection the considerations involved in single-diode detection are fairly simple. Aside from the differences effected by the application of limiting to the carrier, about all that is required is to add sufficient carrier to the carrier of the unfiltered signal, such that the resultant lowering of the percentage of modulation fed to the detector will provide a margin which will allow carrier fading without producing overmodulation. There are limits to the degree of exaltation that should be applied, however.

One factor controlling the degree of carrier exaltation that may be used has to do with the relative amounts of noise contributed to the receiver output by the carrier channel and the unfiltered signal channel. With single-diode detection, each channel contributes to the detected output in accordance with its amplitude at the detector input. Under this condition, if excessive carrier exaltation were applied to the detector, the carrier channel might contribute a major proportion of the noise to the output. This would impair the signal-to-

noise ratio because the output of the signal does not increase as the amplitude of the filtered carrier fed to the detector is increased. Hence, increasing the carrier amplitude increases the noise contributed by the carrier channel, but does not increase the detected signal output. This independence of the signal output and filtered-carrier amplitude is brought about by the fact that the increase in detector current, effected by an increase in filtered-carrier amplitude, is compensated for by a lowering of the effective percentage of modulation fed to the detector.

The degree with which the carrier channel will contribute noise to the receiver output depends upon the degree of carrier exaltation and the difference between the bandwidths of the unfiltered-signal channel and the carrier channel. Since the carrier channel is much narrower than the unfiltered-signal channel, a considerable degree of carrier exaltation is allowable before the carrier-channel noise rises to a level comparable to that from the unfiltered channel. In the case of a typical experimental receiver having an intermediate-frequency channel 10 kilocycles wide, the carrier channel had an equivalent bandwidth of approximately 75 cycles. This is a total-channel-to-carrier-channel bandwidth ratio of 133 times. For equal input carrier-to-noise ratios, the noise at the output of these two channels would be proportional to the square root of the bandwidth ratio, or $\sqrt{133}=11.5=21$ decibels. For this detecting system, there is, therefore, 21 decibels less noise on the carrier channel than on the unfiltered signal channel. Most of this noise is in the low-frequency region where the normal falling off in the response of the audio-frequency amplifier system will effect further attenuation. An approximate evaluation of the amount of attenuation effected in this manner yielded the figure of 5 decibels. This evaluation used the above-mentioned crystal filter with a 75-cycle equivalent bandwidth and an audio system having a response which is down 1 decibel at 50 cycles, 5 decibels at 30 cycles and 12.5 decibels at 20 cycles. The audio attenuation was added to the crystal-filter attenuation to give a reduced equivalent bandwidth from which the relative amount of noise could be evaluated.

Since the carrier-filter noise (for the receiver of this example) is 21 decibels down due to the carrier-filter selectivity, and an additional 5 decibels down due to the audio-frequency attenuation, a carrier exaltation of approximately 26 decibels may be used without bringing the carrier-filter noise above the unfiltered-signal noise. In the practical case, a greater value of exaltation would be usable if desired since the noise contributed by the carrier filter is in the very-low-frequency region where the ear is less sensitive.

The use of carrier limiting with single-diode detection has the advantage that residual amplitude modulation remaining on the filtered carrier is removed. This residual amplitude modulation may take the form of hum and low-frequency sidebands which pass the selectivity of the carrier filter, or it may consist of the amplitude-modulated noise that normally would be contributed by the carrier channel. The effect is to allow the use of a greater degree of carrier exaltation without encountering overemphasis of low-modulation frequencies or contributions of hum and noise from the carrier channel.

It is worthy of note that the carrier limiter is not indispensable with single-diode detection. The fact that the output of such a detector is independent of carrier strength, removes the volume variations, due to carrier fading, from the audio output.

For phase-modulation detection with the single diode, the only change required is a shift in the phase of combination between the unfiltered carrier and the filtered carrier. The phase difference is switched from the zero-degree relation proper for amplitude-modulation detection to a 90- or 270-degree phase difference. The vector diagrams of Fig. 6 show how the resultant E_R, of the

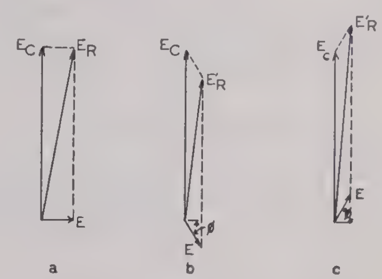

Fig. 6—Vector diagrams for exalted-carrier phase-modulation detection.

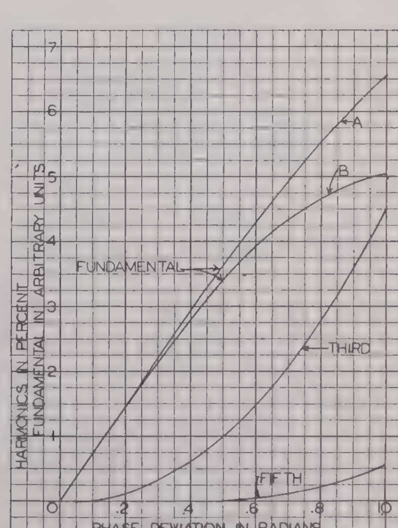

Fig. 7—Fundamental and harmonic output for phase-modulation detection. Curve A = fundamental output for exalted-carrier diode detection with or without carrier limiting, or for multigrid detection with carrier limiting. Curve B = fundamental output for multigrid detection without carrier limiting.

combination of the carrier wave E_c and the unfiltered phase-modulated wave E, varies in amplitude as E varies in phase by the amount ϕ. Fig. 6a is the unmodulated condition and Figs. 6b and 6c show the

effect on the resultant amplitude as the phase is modulated to either side of the unmodulated condition. The amplitude modulation produced on the resultant is detected by the diode detector.

Appendix I gives an analysis of diode phase-modulation detection, in which relations are derived evaluating the variation with phase deviation of the amplitude of the fundamental modulation frequency and the percentage of harmonics. The curves of Fig. 7 have been plotted from these relations. When the carrier exaltation is high, the even harmonics become negligible and the remaining harmonic distortion is substantially all third harmonic. The fundamental output (curve A) deviates somewhat from exact linearity since it is proportional to $J_1(\phi)$.

Considerations with respect to the allowable degree of carrier exaltation remain the same with phase-modulation detection as with amplitude-modulation detection. This is true since the carrier component is able to contribute noise to the output in the same degree regardless of the relative phase relations between the unfiltered signal and the carrier.

Balanced-Diode Detection

Fig. 8 shows the circuit for balanced-diode detection. Diodes C and D have their outputs connected differentially so that their outputs cancel if the amplitude

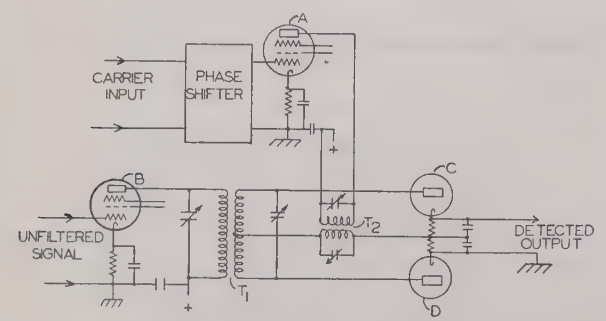

Fig. 8—Balanced-diode detector circuit.

envelopes on their input waves are in phase. Hence, if input is present from either the carrier channel or the unfiltered-signal channel alone, there is no output except that due to unbalance.

For amplitude-modulation detection, the phase of the carrier fed to the balanced detectors is either zero or 180 degrees. This causes the added carrier to aid the signal carrier at the input of one detector and to oppose at the input of the other detector. The diode which is being fed by the opposing combination has the possibility of contributing distortion equivalent to a complete fade-out of the carrier when the amplitude of the added carrier and that of the signal are equal. As a result of this possibility, the degree of carrier exaltation must be kept high to avoid distortion.

Balanced-diode detection has its greatest advantage in phase-modulation reception. As shown in Appendix I, the balanced circuit cancels second-harmonic distortion that is introduced when the carrier fades. The result is the ability to tolerate a greater degree of carrier fading for a given degree of carrier exaltation.

Multigrid Detection

Fig. 9 shows how a multiple-grid converter tube may be used for exalted-carrier detecton.[5] A tube of the 6SA7 type may be used. Capacitor C by-passes carrier and sideband frequencies and C_k and C_s by-pass audio frequencies. Both grids are biased to the linear portion of their characteristics so that there is no detection when a signal is fed to one grid alone. A detector arranged in this way may be looked upon as a linear modulator in which one grid controls the gain of the other grid in the manner of the ordinary amplitude modulator.

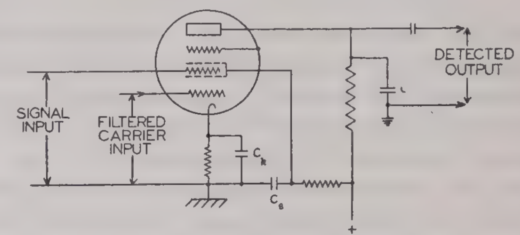

Fig. 9—Multigrid detector circuit.

The multigrid detector does not require the exaltation of the filtered carrier with respect to the signal. Normally, equal voltages are fed to each grid with a zero-degree phase difference for amplitude-modulation detection and 90 degrees for phase-modulation detection. The distortion due to carrier fading is removed inherently in the mechanism of detection. As shown in Appendix II, harmonic distortion is absent in the case of amplitude-modulation detection. For phase-modulation detection the linearity of the fundamental and the amplitude of the harmonics for the case of a limited carrier, are the same as for the case of diode detection when the degree of carrier exaltation is high in the diode detection. Thus, the curves of Fig. 7 portray the harmonic distortion for the case of multigrid detection also.

The output of the multigrid detector is proportional to the product of the amplitudes of the filtered carrier and unfiltered signal fed to the two grids. This makes the detected output proportional to the square of the signal voltage. Such a square-law characteristic accentuates volume variations due to fading unless the automatic-volume-control system is quite efficient. The use of a carrier limiter holds the input to one of the grids constant and thereby changes the characteristic from square-law to a linear relation between unfiltered-signal and output volume. In addition, the limiter removes the amplitude variations brought about by the carrier

[5] Murray G. Crosby, U. S. Patent No. 2,063,588, "Phase or amplitude-modulated wave demodulator," issued December 8, 1936.

amplitude-coefficient $J_0(\phi)$ in phase-modulation reception. This converts the fundamental output linearity from that of curve B to curve A in Fig. 7. The use of a carrier limiter thus has added advantages with this type of detector.

The multigrid detector has the advantage of simplicity in that the two inputs are negative-biased grids. A diode-driver transformer is thus eliminated.

In applying the 6SA7 type of tube to multigrid detection, values of element voltages are chosen which make each grid operate on the linear portions of its characteristic. This may be done by applying an amplitude-modulated signal to a single grid at a time and adjusting for a minimum of detected output. The final adjustment is one in which there is little or no signal output with the modulated signal on one grid, but full output when the signal is applied to both grids. This adjustment may be obtained with a single cathode-resistor bias if the cathode resistor is by-passed for audio frequencies.

Selectivity Characteristics

The use of exalted-carrier detection produces the equivalent of an increase in the selectivity of the receiver. With diode detection, the phenomena of the apparent demodulation of a weak signal by a stronger one[6] is brought into play to reduce the modulation of the undesired signal. Without carrier exaltation, a strong adjacent signal produces an apparent demodulation of the weaker signal so that the weak desired signal may be smothered by the modulation output of the strong undesired signal. The factor by which the modulation output of the weaker desired signal is reduced below the output voltage it develops in the absence of the strong undesired signal[7] is

$$D_1 = (1/2)(E/E_i) \quad \text{for} \quad E_i \geqq 2E \qquad (1)$$

where E/E_i is the ratio of the weak signal to the strong undesired signal.

The original signal-to-interference ratio is equal to the quantity E/E_i. This is multiplied by D_1 due to the demodulation effect of the linear detector. The resulting signal-to-interference ratio without carrier exaltation is thus given by

$$S/I = ED_1/E_i = (1/2)(E/E_i)^2. \qquad (2)$$

When carrier exaltation is applied, the strength of the desired carrier is increased X times with respect to the interfering signal carrier. This invariably has the effect of making the desired signal the strongest, so that the apparent demodulation effect functions to reduce the output of the undesired signal instead of the desired. The demodulation effect in the presence of carrier exaltation will thus be

$$D_2 = (1/2)(E_i/EX). \qquad (3)$$

The signal-to-interference ratio in the presence of carrier exaltation will therefore be

$$S_eI_e = EX/E_iD_2 = 2(EX/E_i)^2. \qquad (4)$$

The signal-to-interference ratio improvement brought about by the use of carrier exaltation is (4) divided by (2) or

$$\text{carrier-exaltation selectivity improvement} = 4X^2. \qquad (5)$$

Equation (5) indicates that the undesired signal is reduced $4X^2$ times from the value it would have in the absence of carrier exaltation. This selectivity improvement is only effective on the modulation component of the undesired signal. The degree of interference from the heterodyne between the desired carrier and the carrier and sidebands of the interference is unchanged. The result is an almost complete removal of voice or program interference from an adjacent channel, but no effect on the whistle due to beats between the carriers or on the "monkey chatter" due to the beats between the desired carrier and the undesired sidebands.

With multigrid detection, the output is proportional to the product of the two signals on the grids. The product of the filtered carrier and the desired signal is the desired signal output, but the product of the filtered carrier and the undesired signal is only the whistle and "monkey chatter" from the beats between the filtered carrier and the carrier and sidebands of the undesired signal. Hence, with a perfect multigrid detector, the modulation component of the undesired signal is completely eliminated. However, practically, the degree of elimination is limited by the degree of linearity that may be obtained on the separate grids of the tube used for the multigrid detector.

In the case of adjacent-channel interference, this selective effect due to carrier exaltation may be classified as a means of substituting audio selectivity for intermediate-frequency selectivity. If sufficient intermediate-frequency selectivity is available properly to select the desired signal and reject the adjacent-channel signal, there is no improvement effected. However, in the case where the undesired signal is in the same channel as the desired, the selectivity effect provides something over and above that obtainable from intermediate-frequency selectivity, in that the modulation component of the undesired signal is rejected.

Diversity Reception

Fig. 10 shows a complete three-receiver diversity receiver for the range of 3-to-24 megacycles. The components of a conventional diversity receiver[8] were used as a basis with added units to combine the three intermediate-frequency outputs and to effect the exalted-carrier detection. The intermediate-frequency outputs

[6] R. T. Beatty, "Apparent demodulation of a weak station by a stronger one," *Wireless Eng. and Exp. Wireless*, vol. 5, p. 300; June, 1928.

[7] E. V. Appleton and D. Boohariwalla, "The mutual interference of wireless signals in simultaneous detection," *Wireless Eng. and Exp. Wireless*, vol. 9, p. 136; March, 1932.

[8] J. B. Moore, "Recent developments in diversity receiving equipment," *RCA Rev.*, vol. 2, p. 94; July, 1937.

were combined to give a single output by means of a three-channel amplifier operated in such a manner that the strongest signal furnished the output. This was done by controlling the gains of the three amplifier channels from the rectified output of the receiver feeding the channel. The three-channel receiver used common first- and second-heterodyne oscillators. The result was a single intermediate-frequency output representative of the strongest signal being received. This output was fed to the exalted-carrier detecting system.

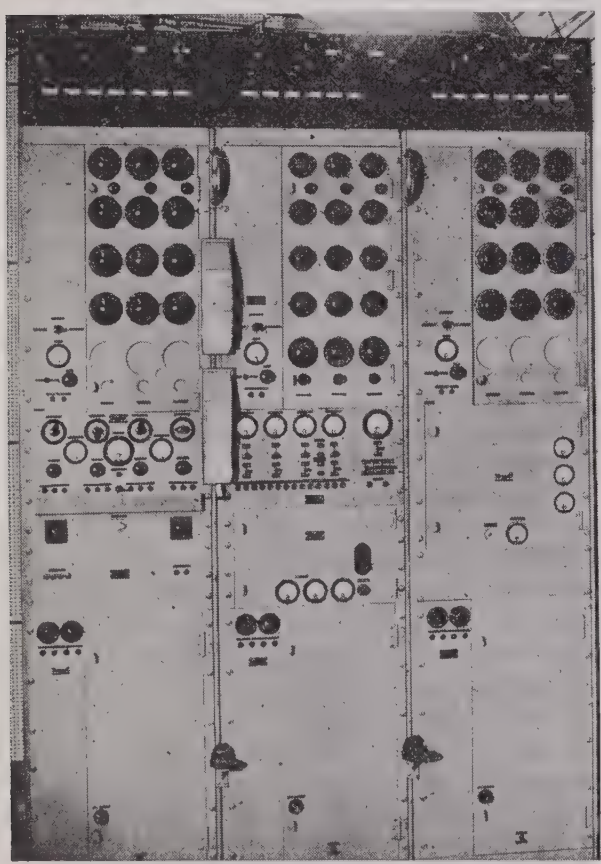

Fig. 10—Experimental exalted-carrier diversity receiver.

In order to obtain some sort of a figure of merit describing the improvement effected by the carrier exaltation, observations were made in which the number of faults marring program reception were counted to determine the relative number of faults per minute. Average figures were obtained from runs totaling 180 minutes in duration for each condition of reception. A fault was counted each time some sort of fading distortion occurred which would be classified as undesirable. The total time was divided into 20-to-50-minute observations during which the condition of reception was switched every 5 minutes. International broadcast signals in the 15-, 11-, and 9-megacycle bands were used, but a major portion of the observations used the 15-megacycle transmission. Table I gives the results of the observations.

TABLE I

Antenna	Unexalted Faults per Minute	Exalted Faults per Minute	Unexalted Faults per Minute / Exalted Faults per Minute
A	3.35	1.17	2.86
B	2.29	0.71	4.2
A and *B* Diversity	2.2	0.43	5.1

It will be noted that antenna *B* shows less faults per minute for both the exalted and unexalted conditions. This effect was noted during the observations and checks were made to insure that it was not due to the receiver. It proved to be a consistent effect that undoubtedly was due to the antennas themselves.

Another outstanding effect is the increase in improvement effected by carrier exaltation when diversity reception is used. This appears to be due to the reduction in depth of carrier fades resulting from the use of diversity. If the carrier fade is deep, it shows up as a fault on the exalted-carrier receiver although it is far less annoying than the distortion produced on the unexalted reception. If the fading is shallow, the distortion is completely removed by exaltation, but a fault would appear on unexalted detection. Thus, diversity reduces the number of deep fades and leaves only the shallow type which may be handled by carrier exaltation.

The fact that a fault occurs during a deep fade on the exalted-carrier reception is not necessarily due to a lack of carrier exaltation. When the carrier fades deeply, the noise level rises since the automatic volume control raises the gain of the receiver inversely with the signal strength. A deep fade may thus produce a burst of noise which would be counted as a fault, regardless of the type of reception.

The application of exalted-carrier detection increases the receiver-output volume variations that normally accompany selective fading. When the carrier fades, the automatic-volume-control system, which is actuated by the carrier energy, operates to increase the gain of the receiver and thereby raises the level of the sidebands fed to the detector. This produces an abnormally high percentage of modulation at the detector input so that the output volume is higher than normal. The usual result is a burst of volume in the receiver output. Such an effect is present with unexalted-carrier detection, but the volume burst is usually a burst consisting of the type of distorted output which is produced when the carrier fades.

One method of removing the volume bursts is by the application of the audio volume-limiting techniques used on the program level fed to broadcast transmitters. This technique was applied on the receiver of Fig. 10. An audio volume limiter was arranged automatically to prevent the receiver output from exceeding a certain maximum level. This arrangement provided a satisfactory solution to the problem.

It should be mentioned that the receiver used in the observations of faults per minute did not have the

benefit of audio volume-limiting equipment. This allowed the occurrence of faults due to volume bursts which would not have been obtained had volume limiting been used. Hence, it may be assumed that even greater improvements than those shown by the table may be expected.

Acknowledgment

The guidance of H. O. Peterson and the assistance of R. E. Schock in the work of this paper are gratefully acknowledged.

Appendix I

Diode Phase-Modulation Detection

The phase-modulated wave is given by

$$e = E \sin (wt + \phi \sin pt) \tag{6}$$

where $w=2\pi F_c$, F_c=carrier frequency, ϕ=peak phase deviation in radians, $p=2\pi F_m$, F_m=modulating frequency.

The resultant wave given by (6) may be resolved to the following carrier and sidebands in a well-known manner.

$$\begin{aligned} e = [&J_0(\phi) \sin wt + J_1(\phi)\{\sin (w+p)t - \sin (w-p)t\} \\ &+J_2(\phi)\{\sin (w+2p)t + \sin (w-2p)t\} \\ &+J_3(\phi)\{\sin (w+3p)t - \sin (w-3p)t\} \\ &+J_n(\phi)\{\sin (w+np)t + (-1)^n \sin(w-np)\}t]. \end{aligned} \tag{7}$$

When the signal given by (6) is subjected to carrier fading, the amplitude of the carrier component varies with respect to the sideband components. This variation may be represented as a subtraction of the carrier component from the total signal wave. Thus

$$e_t = E \sin (wt + \phi \sin pt) - kEJ_0(\phi) \sin wt \tag{8}$$

where the constant k determines the depth of the carrier fading. k is unity for a complete fade-out of the carrier and zero for complete absence of carrier fading.

The resulting carrier component of the fading signal given by (8) is, hence

$$\begin{aligned} e_c &= EJ_0(\phi) \sin wt - kEJ_0(\phi) \sin wt \\ &= EJ_0(\phi)(1 - k) \sin wt. \end{aligned} \tag{9}$$

The carrier filter separates the fading carrier component given by (9) from the sidebands. For diode exalted-carrier reception the filtered carrier is exalted by a factor X and recombined with the fading signal given by (8) to give the following:

$$\begin{aligned} e_t + e_c = E[&\sin (\omega t + \phi \sin pt) \\ &- J_0(\phi)\{k \sin \omega t - X(1 - k) \sin (\omega t + B)\}] \end{aligned} \tag{10}$$

where B is the phase of combination between the exalted carrier and the carrier in the fading signal.

When the degree of carrier exaltation is sufficient to prevent carrier-fading distortion, the quantity k may be neglected in comparison to $X(1-k)$. When this is done, and vector addition is applied to (10), the result is

$$\begin{aligned} e_t + e_c = E\sqrt{1 + X^2J_0^2(\phi)(1 - k)^2 + 2XJ_0(\phi)(1 - k)} \\ \cdot \cos (B - \phi \sin pt) \sin (wt + \alpha) \end{aligned} \tag{11}$$

where α is a variable phase angle dependent on the variables and constants in (9), but of no interest here since the diode is capable of detecting only the amplitude modulation given by the amplitude envelope under the radical in (11).

Equation (11) may be rearranged to give

$$\begin{aligned} e_t + e_c = E\sqrt{1 + X^2(1 - k)^2J_0^2(\phi)} \\ \{1 + A \cos (B - \phi \sin pt)\}^{1/2} \sin (wt + \alpha) \end{aligned} \tag{12}$$

where $A = \{2X(1 - k)J_0(\phi)\}/\{1 + X^2(1 - k)^2J_0^2(\phi)\}$.

The maximum value that A may assume is 1.0. Hence the Binomial Theorem may be applied to (12). When this is done, together, with an application of the formulas for the powers of trigonometric functions, (12) resolves to

$$\begin{aligned} e_t+e_c=E\sqrt{1+X^2(1-k)^2J_0^2(\phi)}\, [&c+c_1 \cos (B-\phi \sin pt) \\ &-C_2 \cos 2(B-\phi \sin pt)+C_3 \cos 3(B-\phi \sin pt) \\ &-C_4 \cos 4(B-\phi \sin pt)+C_5 \cos 5(B-\phi \sin pt) \\ &- \cdots] \sin (wt+\alpha) \end{aligned} \tag{13}$$

where

$$C=1-A^2/16-(15/1024)A^4-105A^6/16{,}384+ \cdots \tag{14}$$

$$C_1=A/2+(3/256)A^3+35A^5/2048+ \cdots \tag{15}$$

$$C_2=A^2/16+5A^4/256+315A^6/32{,}768+ \cdots \tag{16}$$

$$C_3=A^3/256+35A^5/4096+693A^7/131{,}072+ \cdots \tag{17}$$

$$C_4=5A^4/1024+63A^6/16{,}384+12{,}012A^8/4{,}181{,}504 + \cdots \tag{18}$$

$$C_5=7A^5/4096+231A^7/131{,}072+ \cdots \tag{19}$$

$$C_6=21A^6/32{,}768+3632A^8/4{,}181{,}504+ \cdots . \tag{20}$$

For phase-modulation reception, the angle B between the filtered carrier and the signal carrier is made 90 degrees. Substituting this value and applying the Bessel Function expansions to (13) gives

$$\begin{aligned} e_t+e_c=ED\sqrt{1+X^2(1-k)^2J_0^2(\phi)}\, [&1+(D_1/D) \sin pt \\ &+(D_2/D) \sin 2pt+(D_3/D) \sin 3pt \\ &+(D_4/D) \sin 4pt+(D_5/D) \sin 5pt \\ &+ \cdots] \sin (wt+\alpha) \end{aligned} \tag{21}$$

where,

$$D = (C + C_2J_0(2\phi) - C + J_0(4\phi) + \cdots) \tag{22}$$

$$D_1 = (2C_1J_1(\phi) - 2C_3J_1(3\phi) + 2C_5J_1(5\phi) - \cdots) \tag{23}$$

$$D_2 = (2C_2J_2(2\phi) - 2C_4J_2(4\phi) + 2C_6J_2(6\phi) - \cdots) \tag{24}$$

$$D_3 = (2C_1J_3(\phi) - 2C_3J_3(3\phi) + 2C_5J_3(5\phi) - \cdots) \tag{25}$$

$$D_4 = (2C_2J_4(2\phi) - 2C_4J_4(4\phi) + 2C_6J_4(6\phi) - \cdots) \tag{26}$$

$$D_5 = (2C_1J_5(\phi) - 2C_3J_5(3\phi) + 2C_5J_5(5\phi) - \cdots). \tag{27}$$

The quantity in brackets in (21) is the amplitude-modulation envelope of the wave fed to the diode detector. It is known without further investigation that the diode will linearly detect this envelope and produce the sinusoidal terms in the brackets with relative amplitudes in accordance with the product of the amplitude constant of each term and the quantity

$$ED\sqrt{1 + X^2(1 - k)^2J_0^2(\phi)}$$

in front of the bracket. The output of the fundamental is therefore given by the following:

$$e_1 = GED_1\sqrt{1 + X^2(1 - k)^2J_0^2(\phi)} \sin pt \tag{28}$$

where G is the detection constant. The percentage of harmonics is proportional to the ratio between the amplitude coefficient of the harmonic and that of the fundamental, or

$$e_2 \text{ (per cent of } e_1) = 100(D_2/D_1) \sin 2pt \quad (29)$$
$$e_3 \text{ (per cent of } e_1) = 100(D_3/D_1) \sin 3pt \quad (30)$$
$$e_4 \text{ (per cent of } e_1) = 100(D_4/D_1) \sin 4pt \quad (31)$$
$$e_5 \text{ (per cent of } e_1) = 100(D_5/D_1) \sin 5pt. \quad (32)$$

When the degree of carrier exaltation X is high and the carrier fading is such that the quantity $X(1-k)$ is large compared to unity, the series for the factors D, D_1, etc., converge to their first terms, and the even harmonics become insignificant. This simplifies (28) to (31), inclusive, to

$$e_1 = 2GEJ_1(\phi) \sin pt \quad (33)$$
$$e_3 \text{ (per cent of } e_1) = 100(J_3(\phi)/J_1(\phi)) \sin 3pt \quad (34)$$
$$e_5 \text{ (per cent of } e_1) = 100(J_5(\phi)/J_1(\phi)) \sin 5pt. \quad (35)$$

When limiting is applied to the carrier, the effect is to remove the amplitude variations represented by the coefficient $J_0(\phi)$. This makes $J_0(\phi)=1$. The only factor containing the quantity $J_0(\phi)$ is D given by (22). Since the factor D does not enter into the amplitude coefficients of the fundamental or harmonics, it may be concluded that carrier limiting does not effect the linearity of this type of detection.

When the balanced-diode detecting circuit of Fig. 8 is used, the phase of either the filtered carrier or the unfiltered signal is reversed at the input of one of the detectors. This makes the sign of the quantity A negative. If $-A$ is substituted for A in (12) it is seen that the amplitude of the fundamental and odd harmonics is reversed in sign, but the even harmonics are proportional to A^2 and are therefore unaffected. Hence, when the two detector outputs of the balanced detecting system are combined differentially, the fundamental and odd harmonics add, but the even harmonics which predominate at low degrees of carrier exaltation cancel. The fundamental and harmonics are then given by (28), (30), and (32). Thus, by this cancellation of the even harmonics which are encountered with low degrees of carrier exaltation, the balanced detecting system makes possible the use of a lower degree of carrier exaltation for a given amount of distortion.

When carrier limiting is employed with balanced detection, the effect is the same as that with single-diode detection.

Appendix II

Multigrid Amplitude- and Phase-Modulation Detection

The case of the multigrid detection of a wave simultaneously modulated in amplitude and phase has been considered previously.[5] The following treatment considers amplitude- and phase-modulation detection for the case of a fading carrier.

The amplitude-modulated wave is given by

$$e = E(1 + m \sin pt) \sin wt \quad (36)$$

where m is the modulation factor, $w=2\pi F_c$, F_c=carrier frequency, $p=2\pi F_m$, F_m=modulating frequency.

As in Appendix I, the effect of carrier fading on (36) may be represented by

$$e = E(1 + m \sin pt) \sin wt - kE \sin wt \quad (37)$$

which may be rewritten

$$e = E[(1 - k) \sin wt + m \sin pt \sin wt]. \quad (38)$$

The carrier filter rejects the sidebands from (38) and leaves the carrier,

$$e_c = E_c(1 - k) \sin wt. \quad (39)$$

The multigrid detector may be considered as a linear modulator which imparts an envelope $[1+f(t)]$ on the wave being modulated when that wave is fed to one grid and the modulating wave $f(t)$ is fed to the other. If it is assumed that the carrier is the modulating wave, and the unfiltered signal is being modulated, the result is

$$e = E[1 + SE_c(1 - k) \sin wt] \cdot [(1 - k) \sin wt + m \sin pt \sin wt] \quad (40)$$

where the constant S relates the carrier voltage E_c to the degree of amplitude modulation produced by the modulator action of the detector.

Multiplying out (40) gives the detector output, which is

$$e = E[(1 - k) \sin wt + m \sin pt \sin wt + SE_c(1 - k)^2 \sin^2 wt + SE_cm(1 - k) \sin pt \sin^2 wt]. \quad (41)$$

Only the audio-frequency components of (41) are utilized from the detector output. The other terms represent either radio-frequency or direct-current output. The final term in (41) is the only one containing audio frequency. It is

$$e_{af} = SEE_cm(1 - k) \sin pt \sin^2 wt \quad (42)$$

which may be resolved to

$$e_{af} = \{SEE_cm(1 - k)/2\} \sin pt - \{SEE_cm(1 - k)/2\} \sin pt \cos 2wt. \quad (43)$$

The second term of (43) is a radio-frequency output. The first term is the final audio-frequency output, which is

$$e_{af} = (S/2)EE_c(1 - k)m \sin pt. \quad (44)$$

Equation (44) gives the final audio-frequency output of the multigrid detector. The output is proportional to the depth of the originally applied modulation m and to the product of the voltages of the carrier $E_c(1-k)$ and unfiltered signal E fed to the two grids. It will be noted that exaltation of the carrier component is not required to eliminate distortion as was the case in diode phase-modulation detection. The only requirement is that the linear-modulator detector operation be maintained by applying the proper input voltages to the two grids.

The detector output is proportional to the product of the unfiltered signal voltage E and the carrier voltage $E_c(1-k)$. Since E_c is derived from E in the carrier filtering process, the output is proportional to E^2. This

makes the audio output proportional to the square of the carrier voltage. This square-law volume variation may be made linear by the application of limiting to the carrier. Limiting makes $E_c(1-k)$ constant, so that the output is directly proportional to the signal voltage E.

When the phase-modulated wave given by (8) and the filtered carrier given by (9), of Appendix I, are applied to the multigrid detector, the result is

$$e_d = E[1 + SE_cJ_0(\phi)(1 - k)\sin(wt + B)] \cdot [\sin(wt + \phi \sin pt) - kJ_0(\phi)\sin wt] \quad (45)$$

where B is the phase of combination between the filtered carrier and the carrier of the fading signal. Equation (45) multiplies out to

$$e_d = E\sin(wt + \phi\sin pt) - kEJ_0(\phi)\sin wt + SEE_cJ_0(\phi)(1 - k)\sin(wt + \phi\sin pt)\sin(wt + B) - SEE_cJ_0^2(\phi)(1 - k)k\sin wt\sin(wt + B). \quad (46)$$

The only audio-frequency term in (46) is

$$e_{af}=SEE_cJ_0(\phi)(1-k)\sin(wt+\phi\sin pt)\sin(wt+B) \quad (47)$$

which may be rewritten

$$e_{af} = (SEE_cJ_0(\phi)(1 - k)/2)\{\cos(\phi\sin pt - B) - \cos(2wt + \phi\sin pt + B)\} \quad (48)$$

from which the audio-frequency component is

$$e_{af} = (S/2)(EE_cJ_0)(\phi)(1 - k)\cos(\phi\sin pt - B). \quad (49)$$

For phase-modulation reception, the angle B is 90 degrees. Substituting this for B in (49) and applying the Bessel Function expansions gives

$$e_{af} = SEE_cJ_0(\phi)(1 - k)[J_1(\phi)\sin pt + J_3(\phi)\sin 3pt + J_5(\phi)\sin 5pt + \cdots] \quad (50)$$

from which the fundamental-modulation component is

$$e_1 = SEE_cJ_0(\phi)(1 - k)J_1(\phi)\sin pt \quad (51)$$

and the harmonics (as a per cent of the fundamental amplitude)

$$e_3 = 100(J_3(\phi)/J_1(\phi))\sin 3pt \quad (52)$$

$$e_5 = 100(J_5(\phi)/J_1(\phi))\sin 5pt. \quad (53)$$

The harmonic content given by (52) and (53), for the multigrid detector is the same as that given by (34) and (35) for the diode detector for the condition of a high degree of carrier exaltation. The fundamental amplitude, given by (51) is similar to that obtained with amplitude modulation using the multigrid detector, except that the coefficients $J_0(\phi)$ and $J_1(\phi)$ are included. This gives a linearity characteristic in the absence of carrier limiting which is somewhat different from that obtained with diode detection. The presence of the term $J_0(\phi)$ causes the linearity to be somewhat more drooping. When carrier limiting is applied, $J_0(\phi)$ is held constant at a value of unity so that the fundamental output is proportional to $J_1(\phi)$ in the same manner as exalted-carrier diode detection.

Electron-Repulsion Effects in a Klystron*

L. A. WARE†, ASSOCIATE, I.R.E.

Summary—**This paper is an extension of an article by V. Y. Savelyev appearing under the title "On the Theory of the Klystron," in the Journal of Technical Physics (Russian), vol. 10, no. 16; 1940. Appreciation is here expressed for the permission granted to The Institute of Radio Engineers to present portions, modifications or extensions of this material in the pages of its PROCEEDINGS. Savelyev's work is a review of some of the results of Webster, and seems to present a simpler method of development, especially in respect to electron-repulsion effects in the beam. In this article the use of the klystron as an amplifier and as an oscillator, and the optimum operating conditions are considered. The restricting assumption is that the depth of modulation be small, and the time of flight of an electron through the buncher is neglected. It is found that, in the case of an amplifier, the presence of electron-repulsion debunching sets a limit to the upper value of length of drift space which can be used. Thus the gain is limited. In the case of the oscillator operating under optimum conditions the debunching has very little effect.**

I. Introduction

As IS WELL known, the klystron[1-5] is an excellent example of the application of velocity modulation. This method of modulation makes possible the development of both amplifiers and oscillators, operating at ultra-high frequencies, which are not restricted by transit time effects. In fact, the transit time is utilized in order to accomplish the modulation. To anyone beginning a study of the klystron, two questions immediately present themselves. One has to do with the consequences of the assumption that the depth of modulation is very small. The other is concerned with the limitations arising from the mutual repulsion of electrons in the beam causing a debunching effect. This paper is primarily concerned with the second of these. Referring to Fig. 1 the following definitions and assumptions are made. As in Webster's article the subscript 0 applies to the cathode; 1 to the buncher, and 2 to the catcher. The separation between centers of the pairs of grids is taken as S. Capital V's refer to voltages and lower case v's to velocities. The voltage V_1 is

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† State University of Iowa, Iowa City, Iowa.

[1] For related work on the klystron see the following articles: V. P. Gulyaev, "On the theory of the klystron," *Jour. Tech. Phys.* (Russian), vol. 11, pp. 101–105; 1941.

[2] D. L. Webster, "Cathode ray bunching," *Jour. Appl. Phys.*, vol. 10, pp. 501–508; July, 1939.

[3] D. L. Webster, "Theory of klystron oscillations," *Jour. Appl. Phys.*, vol. 10, pp. 864–872; December, 1939.

[4] Arthur E. Harrison, "Graphical methods for analysis of velocity-modulation bunching," PROC. I.R.E., vol. 33, pp. 20–32; January, 1945.

[5] A. E. Harrison, "Klystron technical manual," (bibliography), Sperry Gyroscope Company, Inc., Brooklyn, N. Y., 1944.

assumed to be not more than $0.4V_0$, the ratio V_1/V_0 being defined as the "depth of modulation." It is assumed that the entire change in velocity of the electron caused by V_1 occurs at the point midway between the grids. The electrical quantities are in electrostatic units unless otherwise stated.

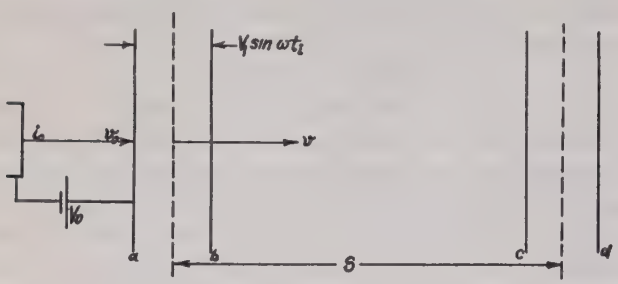

Fig. 1—Basic quantities in a klystron.

In the space S between b and c it is assumed that no field exists except that caused by the electron stream itself.

On the basis of Fig. 1 the following equations can be written

$$v_0 = \sqrt{2eV_0/m} \tag{1}$$

$$v_i = \sqrt{2e/m}\sqrt{V_0 + V_1 \sin \omega t_i} \tag{2}$$

where t_i is the time of passage of the i-th electron through the buncher. From (2)

$$\begin{aligned} v_i &= v_0\sqrt{1 + (V_1/V_0) \sin \omega t_i} \\ &\doteq v_0(1 + (V_1/2V_0) \sin \omega t_i). \end{aligned} \tag{3}$$

The electrons leaving the buncher with velocities given by (3) will form into bunches as they approach the catcher. To what extent this bunching process develops will depend on the value of S, as well as other parameters.

When the i-th electron enters the catcher it gives up an energy

$$\Delta E_i = eV_2 \sin (\omega a_i + \phi) \tag{4}$$

where a_i is the time at which the electron passes the center plane of the catcher and ϕ is a phase constant. Now, if no repulsion effects are assumed along the beam the times a_i and t_i will be related by

$$a_i = t_i + S/v_i = t_i + S/v_0 - (SV_1/2v_0V_0) \sin \omega t_i. \tag{5}$$

The total power, W_2, in the catcher can be obtained as follows. Substitute (5) into (4), and then average for one electron over a period.

$$\Delta E_i = eV_2 \sin (\omega t_i - (\omega SV_1/2v_0V_0) \sin \omega t_i + (\omega S/v_0) + \phi) \tag{6}$$

$$E_{av} = eV_2 \int_0^{2\pi/\omega} \frac{\omega}{2\pi} \sin (Q) dt_i \tag{7}$$

where Q is the argument of the sin function in (6). This is the average energy for one electron over one period. Let this be multiplied by the number of electrons leaving the cathode per second to obtain the power in the catcher, W_2. The number of electrons leaving the cathode per second will be $n = i_0/e$. Then[6]

[6] See Appendix for the integration of (9).

$$W_2 = eV_2 \frac{i_0\omega}{2e\pi} \int_0^{2\pi/\omega} \sin (Q) dt_i \tag{8}$$

$$= \frac{V_2}{2\pi} i_0 \int_0^{2\pi} \sin (Q) d(\omega t_i) \tag{9}$$

$$= i_0 V_2 J_1(S\omega V_1/2v_0V_0) \sin (\phi + S\omega/v_0) \tag{10}$$

where J_1 is the Bessel function of the first order.

The power W_2 will be a maximum if

$$\phi + (S\omega/v_0) = (\pi/2) + 2n\pi \tag{11}$$

and

$$S\omega V_1/2v_0V_0 = 1.84. \tag{12}$$

Accordingly, the maximum power will be

$$W_{2(\max)} = i_0 V_2 J_1(1.84) = 0.58 i_0 V_0 \tag{13}$$

and the efficiency of conversion from direct current to alternating current is, in an ideal case,

$$\eta = W_{2(\max)}/W_0 = 0.58. \tag{14}$$

II. Effect of Electron Repulsion

The above development, arriving at familiar results, serves as a sufficient basis on which to develop the results in the case of electron repulsion along the beam. Strictly speaking, the beam is assumed to have a very great cross section but it will be taken here to be A, of the order of one square centimeter. The electron repulsion will be introduced by substituting for (5) the following expression:

$$a_i = t_i + \int_0^S \frac{1}{v_i} dS' \tag{15}$$

where S' represents a point between b and c in the drift space. This equation allows for a variaton in v_i as a function of S'. Differentiate (15)

$$\frac{da_i}{dt_i} = 1 - \int_0^S \frac{1}{v_i^2} \frac{dv_i}{da_i} \frac{da_i}{dt_i} dS' \tag{16}$$

where dv_i/da_i represents the acceleration of the i-th electron in the field of the space charge, and a_i, here, is the time of passage of the electron through the point S'. This acceleration depends on the field $\mathcal{E}$ as follows:

$$dv_i/da_i = e\mathcal{E}/m \tag{17}$$

and $\mathcal{E}$, in turn, is related to the current as follows: Let i represent the current at point S''. Then $(i - i_0)$ will represent the concentration at the point, and since the excess of electrons, over some average density, is the cause of $\mathcal{E}$,

$$\begin{aligned} \rho &= (i - i_0)/v_iA \\ d\mathcal{E}/dS'' &= -4\pi\rho = -4\pi(i - i_0)/v_iA \end{aligned} \tag{18}$$

$$\mathcal{E} = \frac{-4\pi i_0}{A} \int_0^{S'} \frac{i - i_0}{v_i i_0} dS'' \tag{19}$$

where S'' is the distance to some point in question between $S=0$ and $S=S'$. The law of conservation of charge gives

$$i_0 dt_i = i da_i$$

or

$$i/i_0 = dt_i/da_i. \tag{20}$$

Let this be substituted into (19) and then into (17) and (16).

$$da_i/dt_i - 1 = \int_0^S \frac{1}{v_i^2}\left[\frac{4\pi e i_0}{Am}\int_0^{S'}\frac{1}{v_i}\left(\frac{dt_i}{da_i}-1\right)dS''\right]\frac{da_i}{dt_i}dS'. \quad (21)$$

Now if, in the space $b-c$, only one bunch of electrons is formed or has begun to be formed, then the derivative da_i/dt_i changes very little over the path of integration and can be moved into the square bracket. Also v_i can be assumed equal to v_0 owing to the small value of V_1/V_0. Then (21) can be written

$$\frac{da_i}{dt_i} - 1 = \frac{-4\pi e i_0}{Amv_0^3}\int_0^S\int_0^{S'}\left(\frac{da_i}{dt_i}-1\right)dS''dS'. \quad (22)$$

Now set $$\alpha^2 = 4\pi e i_0/mAv_0^3$$
and $$Y(S) = da_i/dt_i - 1$$
and differentiate (22) twice with respect to S.

$$d^2Y(S)/dS^2 + \alpha^2 Y(S) = 0. \quad (23)$$

The solution of (23) is

$$Y(S) = A \sin \alpha S + B \cos \alpha S. \quad (24)$$

$Y(S)$ must meet certain boundary conditions. At $S=0$, from (16), $$Y(S) = 0, \qquad (S = 0). \quad (25)$$

Also from (16)

$$\frac{dY(S)}{dS} = -\frac{d}{dS}\int_0^S \frac{1}{v_i^2}\frac{dv_i}{dt_i}dS'$$
$$= -(1/v_i^2)(dv_i/dt_i) \text{ (evaluated at } S = 0)$$
$$= -(1/v_i^2)(v_0V_1\omega/2V_0)\cos \omega t_i$$

and since, for this purpose, $v_0 \doteq v_i$

$$dY(S)/dS = -V_1\omega/2v_0V_0 \cos \omega t_i \quad (26)$$

because (16) gives v_i at $S=0$. The functions $Y(S)$ will satisfy conditions (24), (25), and (26) if

$$Y(S) = -(\omega V_1/2v_0V_0)\cos \omega t_i(\sin \alpha S/\alpha) = da_i/dt_i - 1. \quad (27)$$

From (27) one can write the expression for a_i.

$$da_i = dt_i - (\omega V_1/2v_0V_0)(\sin \alpha S/\alpha)\cos \omega t_i dt_i$$

or on integration and applying the condition that an electron passing the buncher at $t_i=0$ must be in the center of a bunch and unaffected by the field $\mathcal{E}$, [i.e., $a_i=t_i+(S/v_0)$], we have

$$a_i = t_i + (S/v_0) - (SV_1/2v_0V_0)(\sin \alpha S/\alpha S)\sin \omega t_i. \quad (28)$$

On comparison of this equation with (5) it will be seen at once that in analogy with (10)

$$W_2 = i_0V_2J_1(\omega SV_1 \sin \alpha S/2v_0V_0\alpha S)\sin(\phi + (S\omega/v_0)). \quad (29)$$

Thus the effect of electron repulsion appears as a correction $(\sin \alpha S)/\alpha S$ applied to the drift distance S. This correction factor is less than unity and means that, in general, repulsion effects tend to decrease the effective value of S. Another way of looking at it is to note that for maximum power, when

$$(\omega V_1S/2v_0V_0)(\sin \alpha S/\alpha S) = 1.84,$$

S must be greater than if the repulsion effects are neglected. In other words, the repulsion retards the bunching process and more space must be provided in which it may occur.

Equation (29) is suitable for both amplifiers and oscillators, and is subject to the condition that V_1/V_0 be small.

III. The Klystron as an Amplifier

In speaking of the amplification of a klystron the effective shunt resistances R of the cavities will be taken to be equal and it will be assumed that the power needed for modulation in the buncher is negligible in comparison to the loss in the buncher. The load on the klystron will be taken to be some effective resistance Z shunted across R.

The power developed in the catcher may then be written

$$W_2 = i_0V_2J_1\left(\frac{\omega SV_1}{2v_0V_0}\frac{\sin \alpha S}{\alpha S}\right) = \frac{V_2^2}{2}\left(\frac{1}{R}+\frac{1}{Z}\right). \quad (30)$$

Dividing this equation by V_1V_2 and rearranging, we obtain

$$\mu = \frac{V_2}{V_1} = \frac{2i_0}{V_1}J_1\left(\frac{\omega SV_1 \sin \alpha S}{2v_0V_0\alpha S}\right)\cdot\frac{RZ}{R+Z}\cdot \quad (31)$$

This equation indicates that μ is dependent in general upon the magnitude of V_1. Thus the amplification will be nonlinear. However, if the argument of the Bessel function is sufficiently small so that the approximation $J_1(x)=x/2$ may be used, then

$$\mu = (i_0\omega S/2v_0V_0)(\sin \alpha S/\alpha S)RZ/(R+Z) \quad (32)$$

which is independent of V_1 thus producing linear amplification. In order to have an error of not more than 5 per cent the argument must meet the following condition:

$$(\omega SV_1/2v_0V_0)(\sin \alpha S/\alpha S) \leqq 0.63. \quad (33)$$

If the load on the klystron is such that $Z=R$ then a simpler form of (32) results.

$$\mu \doteq (i_0\omega SR/4v_0V_0)(\sin \alpha S/\alpha S) \quad (34)$$

and if, in addition, $\alpha S \ll 1$,

$$\mu \doteq i_0\omega SR/4v_0V_0. \quad (34')$$

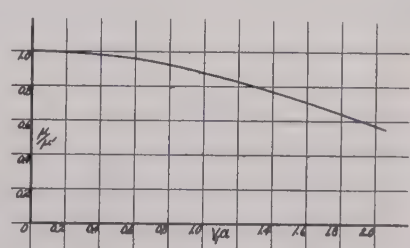

Fig. 2—Variation of klystron amplification with a change in input voltage.

The equation for power amplification is found as follows: the power input is approximately $W_1 = V_1^2/2R$. The output power is determined from (30) as

$$W_s = \frac{V_2^2}{2Z} = \frac{i_0V_2\omega SV_1}{4v_0V_0}\frac{\sin \alpha S}{\alpha S} - \frac{V_2^2}{2R}.$$

The power amplification is then

$$k = \frac{W_s}{W_1} = \frac{i_0 V_2 \omega S R}{2 v_0 V_0 V_1} \frac{\sin \alpha S}{\alpha S} \cdot \frac{V_2^2}{V_1^2}$$
$$= \mu^2 (R/Z) \text{ (using (32)).} \quad (35)$$

From (31) the following equation can be written:

$$\mu/\mu' = (2/V_1 a) J_1(V_1 a) \quad (36)$$

where μ' is the amplification factor at very low values of depth of modulation, and

$$a = (\omega/2 v_0 V_0)(\sin \alpha S/\alpha). \quad (37)$$

Equation (36) is plotted in Fig. 2 and the curve indicates clearly how the gain is dependent upon V_1. The deviation of the ratio from unity reaches 5 per cent at about $aV_1 = 0.63$ as stated above.

IV. Optimum Conditions in Amplification

A brief inspection of (31) shows that the amplification will be maximum with good linearity ($aV_1 \leqq 0.63$) if i_0 is large and V_1 is small. Thus, in the design of a klystron amplifier to operate with a high gain and at a low level of input, it is necessary to use an i_0 as large as practicable and to adjust S so that, for the given input voltage, the argument of the Bessel function is near to the value of 0.63 or less. A numerical example will explain this more

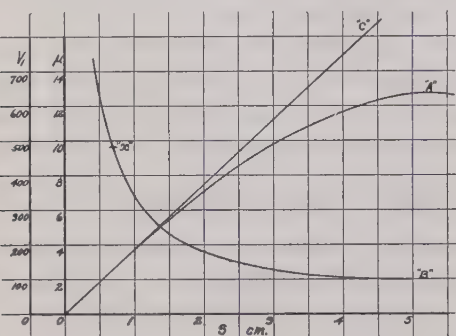

Fig. 3—Curves for a klystron amplifier.
A = gain ratio as a function of S.
B = permissible upper limit of V_1 set by linearity. The point X is the upper limit of V_1 set by the assumption as to depth of modulation.
C = gain ratio as a function of S neglecting debunching.

clearly. Let it be assumed that a klystron is designed to have the following parameters:

$i_0 = 0.020$ ampere $= 0.06 \times 10^9$ electrostatic unit
$V_0 = 1200$ volts
$\omega = 6\pi \times 10^9$
$A = 0.5$ square centimeter
$v_0 = 2.1 \times 10^9$ centimeters per second
$R = 10^5$ ohms
$\alpha = 0.293$

Equation (31) then becomes with $Z = R$,

$$\mu = (2000/V_1) J_1[0.01275 V_1 \sin (0.293S)] \quad (38)$$

where V_1 is in volts. As long as the argument of the Bessel function is less than 0.63 the gain will not vary appreciably with V_1; thus the maximum gain, with this limitation, will occur when $\sin 0.293S = 1$, or $S = 5.36$ centimeters. The upper limit of V_1, at this value of S, will then be given by $0.01275 V_1 = 0.63$, or $V_1 = 49.5$ volts. Up to this limit the gain as a function of S will be given by the curve A in Fig. 3. The permissible upper limit of V_1 to be used with a given value of S is given by curve B. With these curves is plotted also the value of μ as a function of S neglecting the effect of electron repulsion. This is curve C. It clearly indicates that, for any particular value of gain, S must be greater when electron-repulsion effects are taken into consideration.

V. The Klystron as an Oscillator

The amplifier is transformed into an oscillator by providing a feedback connection between the buncher and the catcher. This feedback supplies to the buncher a certain fractional part of the generated voltage in the catcher. Let this be bV_2. Assuming optimum phase conditions which will restrict the value of V_0, the condition for the beginning of self-sustained oscillations can be developed as follows: it will be assumed that the output power is the same as the power loss in the catcher and that the power required for modulation in the buncher can be neglected in comparison with the power loss in the input cavity.

The condition for the beginning of an oscillation will be given by setting the gain from (31) equal to or greater than $1/b$. This is expressed as follows:

$$\frac{2i_0}{V_1} J_1 \left(\frac{\omega S V_1}{2 v_0 V_0} \frac{\sin \alpha S}{\alpha S} \right) \cdot \frac{R}{2 + b^2} \geqq \frac{1}{b} \cdot \quad (39)$$

For initial oscillations this *inequality* must hold for very small values of V_1; i.e.,

$$(i_0 \omega S/2 v_0 V_0)(\sin \alpha S/\alpha S) R/(2 + b^2) > 1/b$$

or $\qquad i_0 > 2 v_0 V v (2 + b^2)/(\omega R b S (\sin \alpha S/\alpha S)). \quad (40)$

The steady-state condition will be indicated by the *equality* (39) which can be expressed as follows:

$$\frac{2i_0}{V_2} J_1 \left(\frac{\omega S b V_2}{2 v_0 V_0} \frac{\sin \alpha S}{\alpha S} \right) \cdot \frac{R}{2 + b^2} = 1. \quad (41)$$

For maximum efficiency and power the argument of the Bessel function must be 1.84, and $V_2 = V_0$. With these introduced into (41) the result is

$$(2 i_0 (0.58)/V_0) R/(2 + b^2) = 1 \quad (42)$$

with a restriction on b that $V_1 \leqq 0.4 V_0$.

Equation (40) provides a lower limit for i_0 and (42) an upper limit so that

$$2 v_0 V_0 (2 + b^2) \alpha/\omega R b \sin \alpha S \leqq i_0 \leqq (2 + b^2) V_0/1.16R \quad (43)$$

subject to the condition that the argument of the Bessel function be 1.84.

In order to investigate the conditions underlying the operation of the oscillator, let (41) be rewritten as follows:

$$J_1((\omega S b V_2/2 v_0 V_0)(\sin \alpha S/\alpha S)) = V_2 (2 + b^2)/(2 i_0 R). \quad (44)$$

If the two members of this equation are thought of as functions of V_2 and plotted separately, it can be said that oscillations will occur if the left member is larger

than the right for small values of V_2, and that the point of operation will be determined by the crossing point of the two curves. The left member is a Bessel curve and the right, a straight line. The diagram which results is similar to the characteristic curves of a shunt direct-current generator. It is immediately noted that the Bessel curve is a function of all the klystron parameters, whereas the slope of the straight line is a function of only b, R, and i_0. If maximum efficiency is to be obtained, then the operating point (the crossing point) should be such that $J_1(x)=0.58$ and $V_2=V_0$ as previously stated. This means that the straight line and Bessel curve both should pass through the point $J_1(x)=0.58$ and $V_2=V_0$. If it be assumed that A, ω, V_0, (v_0), are fixed at values previously given, the above condition can be met by a number of combinations of b, i_0, and S,

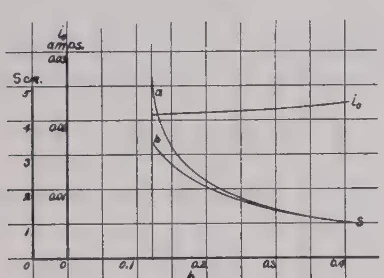

Fig. 4—Combinations of b, i_0, and S which will lead to optimum operation of a klystron oscillator.

as shown in Fig. 4. As indicated, the value of b is limited to values above 0.121 and below 0.4. The latter limit is set in order not to exceed a depth of modulation of 0.4.

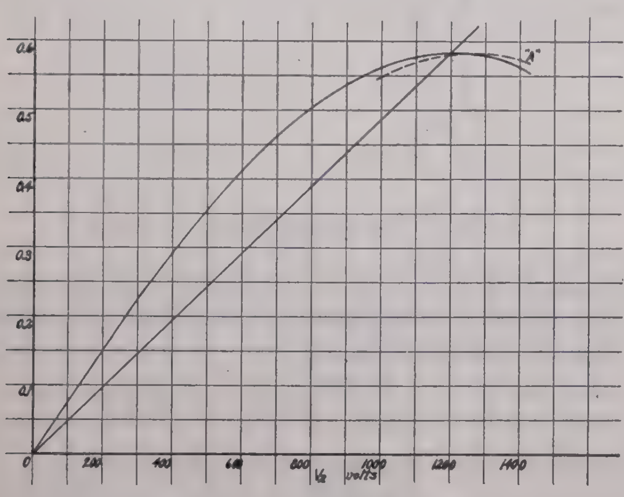

Fig. 5—Characteristic curves for an oscillator based on Fig. 4. The curve A represents the effect of changing S from 2.21 centimeters to 2.05 centimeters.

The lower limit exists because, for values of b less than 0.121, the argument of the Bessel function can not attain the value 1.84 for any value of S. The expressions for S are as follows: for maximum efficiency

$$(\omega b/2v_0\alpha)\sin\alpha S = 1.84$$

or

$$S = (1/\alpha)\sin^{-1}(3.68v_0\alpha/\omega b) \tag{45}$$

or if the debunching effect is neglected

$$S = 3.68v_0/\omega b. \tag{45'}$$

The curves corresponding to (45) and (45′) are shown as a and b respectively in Fig. 4. It is noted that for low values of b the debunching effect is pronounced and, as in the case of the amplifier, it makes necessary a larger value of S for given conditions of operation. Any combination of values of b, i_0, S given by Fig. 4 will produce the same characteristic curves as shown in Fig. 5. For these conditions the output power is 7.2 watts.

It is of interest to determine the effect, upon the operation of a klystron oscillator, brought about by changing various parameters individually. For this purpose let the characteristic curves be plotted for the following parameters:

$$\begin{aligned}
b &= 0.4\\
i_0 &= 0.02 \text{ ampere}\\
\omega &= 6\pi \times 10^9 \text{ radians per second}\\
V_0 &= 1200 \text{ volts}\\
S &= 2.0 \text{ centimeters}\\
v_0 &= 2.1 \times 10^9 \text{ centimeters per second}\\
A &= 0.5 \text{ square centimeter}\\
R &= 10^5 \text{ ohms}
\end{aligned}$$

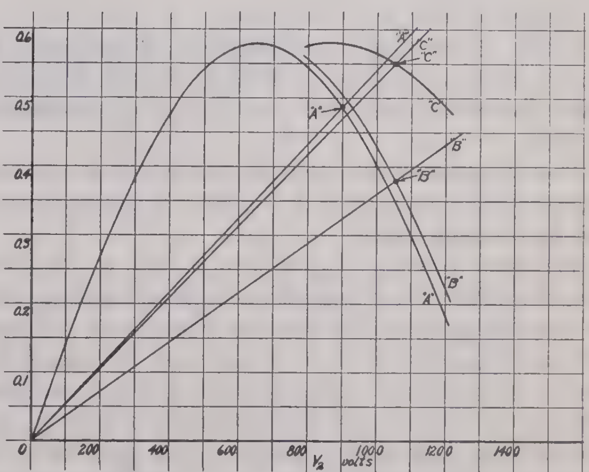

Fig. 6—Characteristic curves for an oscillator with arbitrary parameters.

The curves are plotted as A in Fig. 6, and they indicate that the power output will be 4.05 watts and the efficiency of conversion will be 16.9 per cent. The variations will be considered separately as follows:

A. Variation of S

The parameter S does not affect the slope of the straight line. A decrease in S is seen, by reference to (44), to move the peak of the Bessel curve to the right. This will gradually raise the output power until the peak of the curve cuts the straight line. At this point the power output will be 5.82 watts and the efficiency, 24.2 per cent. A further decrease of S will again lower the power.

B. Variation of i_0

The parameter i_0 primarily affects the straight line and has little effect on the Bessel curve through the debunching factor. An increase in i_0 will decrease the slope of the straight line and thus will increase the power output. A minor effect is that α will increase slightly, causing the debunching factor to decrease. This results in moving the Bessel curve slightly to the right which also increases the power output to a small extent. Curves B for $i_0=0.03$ ampere are plotted in Fig. 6. It is seen that the power has become 5.55 watts and the efficiency has decreased to 15.4 per cent.

C. Variation of b

If b is the only variable and small in value ($\leqq 0.4$), it is seen that it will not greatly affect the straight line but enters into the Bessel curve in the same way as V_2. Thus a change of b from 0.4 to 0.3, as an illustration, should move the Bessel curve to the right and will decrease somewhat the slope of the straight line. The combination of these two effects is to increase the power. Curves C for this case are plotted in Fig. 6. The power increases to 5.55 watts and the efficiency becomes 23.1 per cent.

D. Neglect of Debunching Factor

If the debunching is neglected, the Bessel curve becomes displaced slightly to the left. In the illustration given this would result in a calculated power which is too low.

The analysis for the condition where the straight line cuts the curve to the left of the peak is similar and will not be taken up here.

The effect of changes of parameters on the condition shown in Fig. 5 is of more interest. Here, optimum conditions are assumed for maximum power of 7.2 watts and maximum efficiency of 30 per cent. It is easily seen that a slight change in the position of the Bessel curve will have practically no effect on the power output. Thus a change in S, or what amounts to the same thing, the neglect of the debunching factor, will not appreciably affect the output. For a value of $b=0.2$ the neglect of the debunching factor makes a change in S from 2.21 centimeters to 2.05 centimeters. A short section of the new curve for $S=2.05$ centimeters is shown and it is obvious that practically no change in power occurs. For larger values of b the error is still less.

VI. Conclusion

It may be said in conclusion that electron-repulsion effects are important in the operation of the klystron as an amplifier. Whereas the simplified theory of the amplifier indicates that, with very low input voltage, a very high gain may be obtained by indefinitely increasing S, the actual state of affairs is that debunching sets a definite limit on the amount of gain obtainable. In Fig. 3, for instance, an increase of S beyond about 5.3 centimeters causes a decrease in gain, and the maximum gain obtainable is about 12.7 for the parameters used. This gain may be increased by increasing i_0 but it must be noted that an increase in i_0, everything else remaining fixed, will decrease the debunching factor.

In the operation of a klystron as an oscillator it is found that, generally speaking, very much lower values of S are used. It appears thus that the debunching effect would not be so detrimental. Calculations show that this effect may be neglected as far as maximum power and efficiency are concerned. With increased current, however, this effect may become important.

This brief analysis is deficient because the depth of modulation has been assumed small ($\leqq 0.4$). In oscillators, this may easily be in error. Further, the effect of the finite time of flight of electrons through the buncher grids has been neglected.

Appendix

Integration of (9)

Let the expression to be integrated be written as

$$\int_0^{2\pi} \sin(x + A\sin x + B)dx \qquad (46)$$

where A, and B are defined by (6). Let (46) be expanded into

$$\sin B\int_0^{2\pi}\cos(x + A\sin x)dx + \cos B\int_0^{2\pi}\sin(x + A\sin x)dx. \qquad (47)$$

Let $\phi(A)=\int_0^{2\pi}\sin(x+A\sin x)dx,$

and $x=2\pi-x',\quad x'=2\pi-x.$

$$\text{Then } \phi(A)=\int_{2\pi}^{0}\sin[(2\pi-x')+A\sin(2\pi-x')](-dx')$$

$$=\int_0^{2\pi}\sin[(2\pi-x)+A\sin(2\pi-x)]dx$$

$$=\int_0^{2\pi}\sin(2\pi-x-A\sin x)dx$$

$$=-\int_0^{2\pi}\sin(x+A\sin x)dx=-\phi(A).$$

Therefore $\phi(A)=0$ which eliminates the right-hand term of (47). The left-hand part is evaluated as follows using an equation from the theory of Bessel functions.[7]

$$J_n(z) = \frac{1}{2\pi}\int_0^{2\pi}\cos(n\theta - z\sin\theta)d\theta$$

Identifying $z=-A$, $n=1$, and $\theta=x$, we have

$$J_1(-A) = \frac{1}{2\pi}\int_0^{2\pi}\cos(x + A\sin x)dx.$$

$$\text{Therefore } \sin B\int_0^{2\pi}\cos(x+A\sin x)dx=-2\pi\sin B(J_1(A))$$

$$\text{and } \int_0^{2\pi}\sin(x+A\sin x+B)=-2\pi\sin BJ_1(A)$$

which applied to (9) will yield (10).

[7] G. N. Watson, "Treatise on the Theory of Bessel Functions," second edition, Cambridge University Press, New York, N. Y., 1944, p. 19.

Extending the Frequency Range of the Phase-Shift Oscillator*

RODNEY W. JOHNSON†, MEMBER, I.R.E.

Summary—**Methods for facilitating the design of phase-shift oscillators at extreme frequencies of a fraction of a cycle to a few megacycles are discussed. It is shown that, if the input impedance of each section of a lumped resistance-capacitance network is made K times that of the previous section, the gain required for oscillation can be reduced to a theoretical minimum of 8 for a three-section network with K high, as against 29 for a K of unity. A new circuit element[1] called the "resistance-capacitance transmission line," consisting of a resistance covered by a well-insulated and grounded metal surface, is introduced and is analyzed as a phase-shifting network to give reliable operation in a phase-shift oscillator at frequencies up to a few megacycles. Curves are presented to facilitate the design of the latter type of oscillator. Various configurations of the resistance-capacitance transmission line are discussed and experimental results are presented.**

Introduction

THE PHASE-shift oscillator has appeared frequently in the literature.[2–7] The usual arrangements of the phase-shifting network using lumped constants become either unduly large at the very low frequencies or too small at the high frequencies. In the latter case, stray and tube capacitances limit the maximum frequency obtainable in practical circuits to the upper audio or lower radio-frequency range.

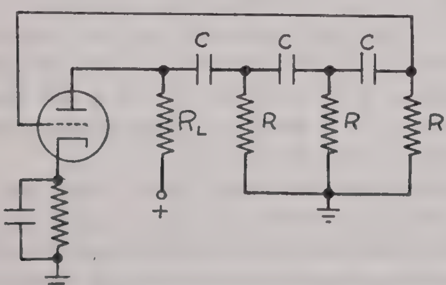

Fig. 1—Conventional phase-shift oscillator.

For example, in the three equal-section resistance-capacitance network of Fig. 1, the minimum gain necessary for oscillation has been shown[2] to be 29. An inspection of the frequency equations of Ginzton and Hollingsworth[2] shows that the frequency range is limited in practical circuits to from perhaps 30 cycles or so to around 50 kilocycles, both because of circuit components and because of loss of gain at extreme ends of the audio range.

It is the purpose of this paper to describe two types of 180-degree resistance-capacitance networks that will extend the useful frequency range to as low as a fraction of a cycle and as high as a few megacycles.

I. Low-Frequency Circuits

Sometimes it becomes necessary to build sinusoidal oscillators of low frequency (say one cycle or so) having considerable power output, such as for timing waves on magnetic oscillographs, recording milliammeters, etc. A gain of 29 under these conditions is difficult to obtain using power tubes. The four-section network, of course, requires a gain[2] of only about 6, but is cumbersome at such a low frequency. The network shown in Fig. 2, however, requires a gain approaching a minimum of 8 at K infinite, and 29 at $K=1$. The circuit of Fig. 2 is of the same type as that previously mentioned, except that each resistance-capacitance section is made to have an input impedance K times that of the previous stage, so that if K is high, the loading due to successive stages is

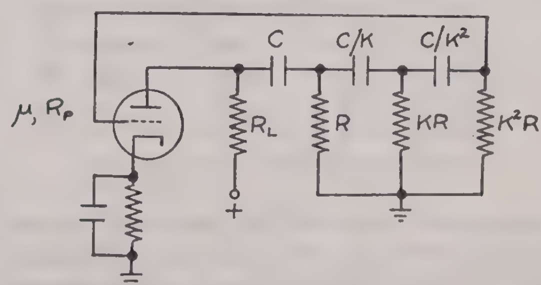

Fig. 2—Revised phase-shift oscillator.

less and the voltage attenuation through the network is consequently reduced. In such a circuit, the voltage attenuation per stage, if K is high, will be very nearly $1/\cos\theta$, where θ is the angle of phase shift for the stage. If the resistance-capacitance products are held constant, then there is a phase shift of 60 degrees per stage, or a total attenuation of $1/\cos^3 60$ degrees$=8$. For a four-stage network, the minimum voltage attenuation would be $1/\cos^4 45$ degrees$=4$. This relationship is strictly true only if both K and the ratio R/R_1 as defined by Fig. 4 are infinite, as is evident from (5).

The circuit of Fig. 3 is equivalent to that of Fig. 2. In Fig. 3, the application of Thevenin's theorem leads to Fig. 4. Fig. 5 is the general equivalent to this type of circuit, and its solution enables us to substitute specific values for any particular case and analyze the result.

The solution of Fig. 5 by Kirchhoff's laws gives the

* Decimal classification: R355.9. Original manuscript received by the Institute, November 20, 1944.
† 2713 Ridge Road, Berkeley 4, California.
[1] C. J. Van Loon, U. S. Patent No. 2,111,710.
[2] E. L. Ginzton and L. M. Hollingsworth, "Phase-shift oscillators," Proc. I.R.E., vol. 29, pp. 43–49; February, 1941.
[3] W. G. Shepherd and R. O. Wise, "Variable-frequency bridge-type frequency-stabilized oscillators," Proc. I.R.E., vol. 31, pp. 256–268; June, 1943.
[4] L. A. Meacham, "The bridge-stabilized oscillator," Proc. I.R.E., vol. 26, pp. 1278–1294; October, 1938.
[5] H. J. Reich, "A low-distortion audio-frequency oscillator," Proc. I.R.E., vol. 25, pp. 1387–1398; November, 1937.
[6] J. H. Newitt, "RC oscillator performance," *Electronics*, vol. 17, pp. 126–129; March, 1944.
[7] W. W. Kunde, "Phase shift oscillator design charts," *Electronics*, vol. 16, pp. 132–133; November, 1943.

ratio of input-to-output voltage as

$$\frac{E}{e}=\left[1+\frac{Z_1Z_5}{Z_4Z_6}+\frac{Z_3Z_5}{Z_4Z_6}+\frac{Z_1Z_3}{Z_2Z_6}+\frac{Z_1Z_5}{Z_2Z_6}+\frac{Z_1Z_3}{Z_2Z_4}\right]+\left[\frac{Z_1}{Z_6}+\frac{Z_3}{Z_6}+\frac{Z_5}{Z_6}+\frac{Z_1}{Z_4}+\frac{Z_3}{Z_4}+\frac{Z_1}{Z_2}+\frac{Z_1Z_3Z_5}{Z_2Z_4Z_6}\right]. \quad (1)$$

If, in (1), Z_1, Z_3, and Z_5 are pure capacitive reactances and Z_2, Z_4, and Z_6 are pure resistances, or vice versa, the first bracketed term will be real and the last bracketed term will be imaginary. Furthermore, all the Z terms

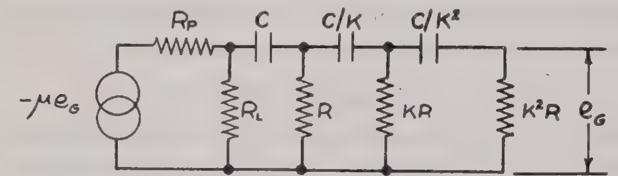

Fig. 3—Equivalent circuit of Fig. 2.

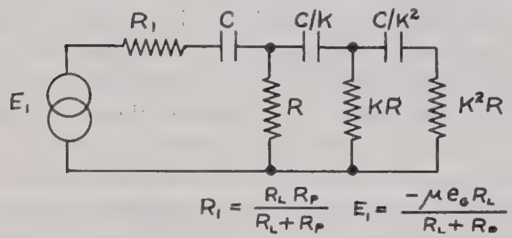

Fig. 4—Equivalent circuit of Fig. 3.

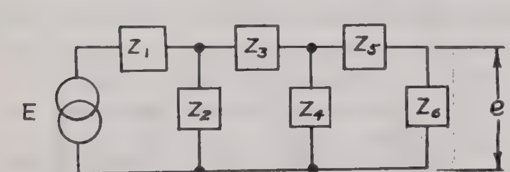

Fig. 5—General phase-shifting network.

will be negative except the very last one. Equating the imaginary series to zero will yield the conditions necessary for 180-degree phase shift, and substitution of this value of R/X or X/R into the first series will give the ratio of input-to-output voltage at this particular frequency, which ratio will be negative, denoting 180-degree phase shift.

When the corresponding values for the Z's from Fig. 4 are substituted into (1), and terms collected,

$$\frac{E}{e}=\left[1-\frac{X^2}{R^2}\left(3+\frac{2}{K}+\frac{R_1}{R}\right)+\frac{R_1}{R}\left(1+\frac{1}{K}+\frac{1}{K^2}\right)\right]-j\frac{X}{R}\left[3+\frac{2}{K}+\frac{1}{K^2}+2\frac{R_1}{R}\left(\frac{1}{K}+1\right)-\frac{X^2}{R^2}\right]. \quad (2)$$

Equating the j terms to zero gives

$$X/R = \sqrt{3 + 2/K + 1/K^2 + (2R_1/R)(1/K + 1)} \quad (3)$$

from which

$$f = 1/2\pi RC\sqrt{3 + 2/K + 1/K^2 + (2R_1/R)(1/K + 1)}. \quad (4)$$

Substitution of (3) into (2) gives the gain which the tube must supply in order to oscillate at a frequency given by (4).

$$\frac{E}{e}=-8-\frac{R_1}{R}\left[\frac{11}{K}+\frac{4}{K^2}+8\right]-\left(\frac{R_1}{R}\right)^2\left(\frac{2}{K}+2\right)-\frac{12}{K}-\frac{7}{K^2}-\frac{2}{K^3}. \quad (5)$$

Setting $K=1$ in the above equations gives the results derived by Ginzton and Hollingsworth[2] and deriving in a similar manner[2] a coefficient of frequency stability, which, if small, is a measure of good frequency stability,

$$k_1 = R_1/[2R_1 + 2R + R(K^2 + 1)/(K^2 + K)] \quad (6)$$

where k_1 is defined as the ratio of a percentage change in frequency to the percentage change in R_1 causing it. Since the quantity $(K^2+1)/(K^2+K)$ is only slightly less than unity for values of K between 1 and infinity, the frequency stability is affected very little by the circuit modification under discussion.

For a practical design using a power pentode such as a 6V6 with $R_p=52{,}000$ ohms, $g_m=4100$, $R_L=5000$ ohms, one could expect a maximum gain from the tube of around 19, which is lower than the required 29 if $K=1$. The maximum grid resistance should not exceed 0.5 megohm and therefore $K^2R<0.5$ megohm. The value of K is determined by this limitation and by the fact that R should be large compared to R_1 for best operation. But to use a lower gain tube, K should be 5 or more. A good compromise indicates $R_L=6000$ ohms, $R=20{,}000$ ohms, and $K=5$. The effective load resistance presented to the tube is around 5000 ohms and the gain should be around 19. Substituting the values $R_1/R=0.27$ and $K=5$ in (5) gives a required gain of 13.7. From (4) the frequency is $3.94/C$ cycles where C is in microfarads.

At audio frequencies much higher than 1000 cycles, the capacitors toward the grid end of the network become rather small and stray circuit capacitance results in unsatisfactory operation; so for higher frequencies we shall now discuss a somewhat different type of resistance-capacitance network.

II. High-Frequency Circuits

The operation of the phase-shift oscillator in the lower radio-frequency range can be made possible by using the equivalent of a resistance-capacitance transmission line[1] which can be very easily constructed by wrapping a metallic foil, with appropriate insulation, around an ordinary resistor and grounding the foil. A total capacitance of around 20 to 50 micromicrofarads can be obtained in this way, which gives very satisfactory results as a 180-degree network when the resistance is connected between grid and plate (with appropriate blocking of direct current) as shown in Fig. 6.

The analysis of this circuit element, which we shall call the resistance-capacitance transmission line, can be carried out using the conventional transmission-line equations. The circuit to be analyzed is shown in Fig.

7. The series inductance and shunt conductance are neglected because they are of very small magnitude compared to the series resistance and shunt capacitance at the frequencies for which this circuit can be used. The effect of series inductance will be shown later.

We shall first analyze the circuit of Fig. 7 to obtain the general equations of voltage and current, and then we shall consider three special cases: the case for open circuit, the case for termination in a capacitance at the grid end, and the case for termination in characteristic impedance.

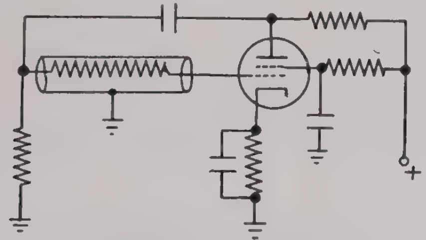

Fig. 6—Circuit of a radio-frequency phase-shift oscillator using the resistance-capacitance transmission line.

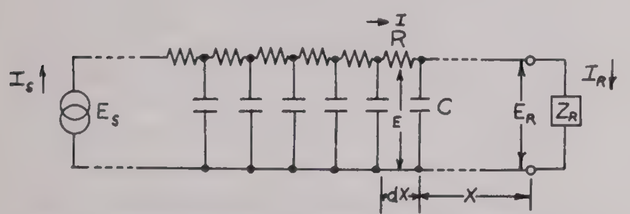

Fig. 7—Equivalent circuit for the resistance-capacitance transmission line.

Referring to Fig. 7, we can write

$$dE/dx = IR, \qquad dI/dx = EY \tag{7}$$

where $Y=j\omega C$.

The solutions of (7) are the familiar transmission line equations

$$E = E_r \cosh \sqrt{RY}\, x + I_r\sqrt{R/Y} \sinh \sqrt{RY}\, x \tag{8}$$

$$I = E_r\sqrt{Y/R} \sinh \sqrt{RY}\, x + I_r \cosh \sqrt{RY}\, x. \tag{9}$$

Equations (8) and (9) will form the basis for investigation of the properties of the resistance-capacitance transmission line for several terminating conditions.

A. *Open Circuit*

1. *Frequency and Gain:* When the output terminals of the network of Fig. 7 are open circuited, $I_r=0$ and (8) reduces to

$$E/E_r = \cosh \sqrt{RY}\, x. \tag{10}$$

Letting

$$\sqrt{RY} = \alpha + j\beta \tag{11}$$

we have, for (10)

$$E/E_r = \cosh (\alpha + j\beta)x$$

$$= \cosh \alpha x \cos \beta x + j \sinh \alpha x \sin \beta x. \tag{12}$$

If $x=l$, the length of the line, $E=E_s$. If we wish 180-degree phase shift to exist between E_r and E_s, the j term of (12) must be zero. Then

$$\sinh \alpha l \sin \beta l = 0 \tag{13}$$

for which case

$$E_s/E_r = \cosh \alpha l \cos \beta l. \tag{14}$$

From (11)

$$\sqrt{RY}=\alpha+j\beta=\sqrt{R\cdot j\omega C}=\sqrt{R\omega C/2}+j\sqrt{R\omega C/2}$$

and

$$\alpha=\beta=\sqrt{R\omega C/2}=\sqrt{\pi f RC}. \tag{15}$$

Now for (13) to be satisfied, either $\sinh \alpha l$ or $\sin \beta l$ must be zero. Since $\sinh \alpha l=0$ only when $\alpha l=0$, and αl cannot be zero for any usable case; then it follows that $\sin \beta l$ must be zero, which is true for $\beta l=m\pi$ (m an integer). Therefore

$$\beta l = \sqrt{\pi f RC}\, l = m\pi. \tag{16}$$

R and C are the constants *per unit length* of the line. If R_T and C_T are the *total* constants of the line (R and C times the length, respectively) then (16) becomes

$$\sqrt{\pi f R_T C_T} = m\pi$$

from which

$$f = m^2\pi/R_T C_T. \tag{17}$$

Substituting (16) into (12) we obtain

$$E_s/E_r = \cosh m\pi \cos m\pi. \tag{18}$$

Equation (18) is positive (0-degree phase shift) for even values of m and negative (180-degree phase shift) for odd values of m. Furthermore, E_s/E_r is the gain the tube must supply in order to oscillate, and for this to be a minimum, m must also be a minimum since $\cosh m\pi$ always increases with increasing m. The least odd positive value of m is unity, and therefore the frequency for minimum required gain at 180-degree phase shift in the resistance-capacitance transmission line is

$$f = \pi/R_T C_T \tag{19}$$

and the required gain is

$$E_s/E_r = -\cosh \pi = -11.60. \tag{20}$$

These are the equations for frequency and gain when the resistance-capacitance transmission line is used in a phase-shift oscillator, *neglecting tube-input capacitance.* The actual case will be investigated later.

2. *Input Impedance:* The input impedance is important, as it is in parallel with the plate and load resistances of the tube, and must be kept high if stable operation is to be maintained. For the open-circuited case at $x=l$, (8) divided by (9) is

$$Z_s = E_s/I_s = \sqrt{R/Y}\,(\cosh \sqrt{RY}\, l)/(\sinh \sqrt{RY}\, l).$$

Or, since $\sqrt{RY}=\alpha+j\beta$ (11)

$$Z_s = \sqrt{\frac{R}{Y}}\ \frac{\cosh \alpha l \cos \beta l + j \sinh \alpha l \sin \beta l}{\sinh \alpha l \cos \beta l + j \cosh \alpha l \sin \beta l}.$$

And for the previously derived case of $\alpha l=\beta l=m\pi$,

$$Z_s = \sqrt{R/Y} \coth m\pi. \tag{21}$$

But

$$\sqrt{\frac{R}{Y}} = \frac{R}{\sqrt{RY}} = \frac{R_T}{l\sqrt{RY}} = \frac{R_T}{\alpha l + j\beta l} = \frac{R_T}{2m\pi} - j\,\frac{R_T}{2m\pi} \tag{22}$$

or, at $m=1$ we have

$$Z_s = \left(\frac{R_T}{2\pi} - j\,\frac{R_T}{2\pi}\right)\coth\pi \cong \frac{R_T}{\sqrt{2}\,\pi}\underline{\big|-45^\circ} \qquad (23)$$

since $\coth\pi$ is very closely unity.

This input impedance looks like a resistance R_T/π in parallel with a capacitive reactance R_T/π. It should be obvious that R_T/π should therefore be at least ten times the output resistance and reactance of the tube for best operation.

B. *Capacitance Termination*

1. *Frequency and Gain:* Let us now consider the actual case where the line is terminated by a capacitance C_G which is the input capacitance of the tube plus strays. From (8)

$$E_s/E_r = \cosh\sqrt{RY}\,l + \sqrt{(R/Y)}/Z_r \sinh\sqrt{RY}\,l. \qquad (24)$$

Letting $\sqrt{(R/Y)}/Z_r = A + jB$ and $\sqrt{RY} = \alpha + j\beta$, (24) becomes

$$\begin{aligned} E_s/E_r &= \cosh(\alpha + j\beta)l + (A + jB)\sinh(\alpha + j\beta)l \\ &= \cosh\alpha l\cosh\beta l + A\sinh\alpha l\cos\beta l \\ &\quad - B\cosh\alpha l\sin\beta l \\ &\quad + j(\sinh\alpha l\sin\beta l + A\cosh\alpha l\sin\beta l \\ &\quad + B\sinh\alpha l\cos\beta l). \end{aligned} \qquad (25)$$

For 180-degree phase shift the j term of (25) must be zero. Since

$$\begin{aligned} A + jB &= \sqrt{(R/Y)}/Z_r = j\omega C_G\sqrt{R/j\omega C} \\ &= \omega C_G\sqrt{R/2\omega C} + j\omega C_G\sqrt{R/2\omega C} \end{aligned}$$

or

$$A = B = (C_G/C_T)\sqrt{\pi f R_T C_T}$$

and since αl and βl are the same as before, the condition for 180-degree phase shift becomes

$$\begin{aligned} &\tanh\sqrt{\pi f R_T C_T}\tan\sqrt{\pi f R_T C_T} \\ &\quad + (C_G/C_T)\sqrt{\pi f R_T C_T}\,(\tan\sqrt{\pi f R_T C_T} \\ &\quad + \tanh\sqrt{\pi f R_T C_T}) = 0. \end{aligned} \qquad (26)$$

Equation (26) cannot be solved explicitly for fR_TC_T, but a sketch of the equation shows that for C_G/C_T and fR_TC_T to be positive, the useful curve lies in the region $3\pi/4$ to π, $7\pi/4$ to 2π, etc. Furthermore, a later equation shows that E_s/E_r for this case becomes too large to be useful for any region but $3\pi/4$ to π. Over this region, $\tanh\sqrt{\pi f R_T C_T}$ is unity to a close approximation and (26) becomes

$$\begin{aligned} &\tan\sqrt{\pi f R_T C_T} \\ &\quad + (C_G/C_T)\sqrt{\pi f R_T C_T}\,(\tan\sqrt{\pi f R_T C_T} + 1) = 0. \end{aligned} \qquad (27)$$

A curve of C_G/C_T versus fR_TC_T is given in Fig. 8. From this curve, knowing the value of C_G/C_T, one can readily determine the value of fR_TC_T which will satisfy (27) and very closely satisfy (26). Inspection of Fig. 8 shows that for $C_G/C_T = 0$ we obtain a value of $fR_TC_T = \pi$ that checks with that given by (19) for the open-circuited case ($C_G = 0$).

With (26) satisfied, we have left only the real term of (25). This is similarly transcendental and cannot be solved explicitly. By taking values of fR_TC_T from Fig. 8 for various values of C_G/C_T, however, and substituting them into the remainder of (25), we arrive at Fig. 9,

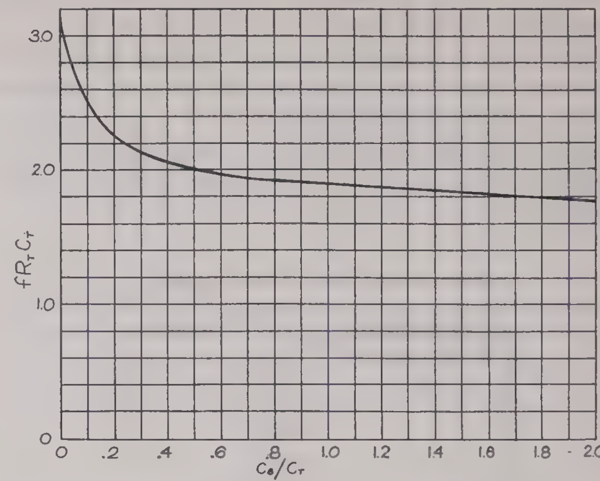

Fig. 8—Curve of frequency for 180-degree phase shift with resistance-capacitance transmission line.

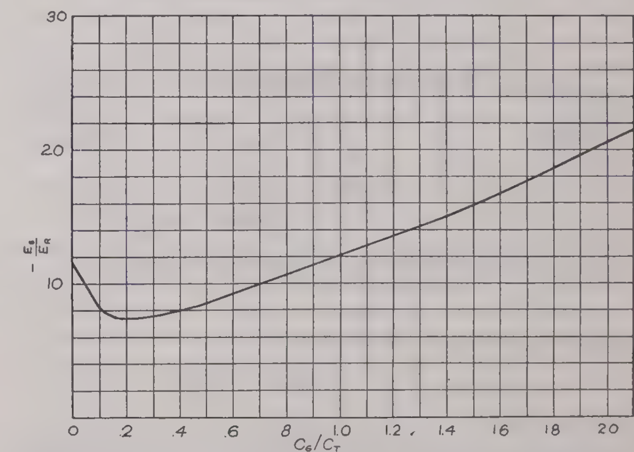

Fig. 9—Curve of voltage attenuation at 180-degree phase shift in the resistance-capacitance transmission line.

which is a curve showing the ratio E_s/E_r as a function of C_G/C_T, for the condition where the phase shift through the resistance-capacitance transmission line is 180 degrees.

2. *Input Impedance:* Returning to (8) and (9), we can find the input impedance in a similar manner to the open-circuited case by dividing one equation by the other. This gives, letting $Z_0 = \sqrt{R/Y}$,

$$\frac{Z_s}{Z_0} = \frac{1 + (Z_0/Z_r)\tanh(\alpha + j\beta)l}{Z_0/Z_r + \tanh(\alpha + j\beta)l}. \qquad (28)$$

Since we have already concluded that $\tanh\alpha l = \tanh\beta l \cong 1$ for all values of αl considered in this analysis; then it is evident that $\tanh(\alpha + j\beta)l \cong 1$ also. Therefore $Z_s = Z_0 = \sqrt{R/Y}$ to within 2 per cent for all values of C_G/C_T considered. The equivalent parallel resistance and reactance components of this impedance are

$$X_s/R_T = -1/\sqrt{\pi f R_T C_T} \qquad (29)$$

$$R_s/R_T = 1/\sqrt{\pi f R_T C_T}. \qquad (30)$$

Taking values of fR_TC_T from Fig. 8 for various values of C_G/C_T, we arrive at Fig. 10, which is a curve of

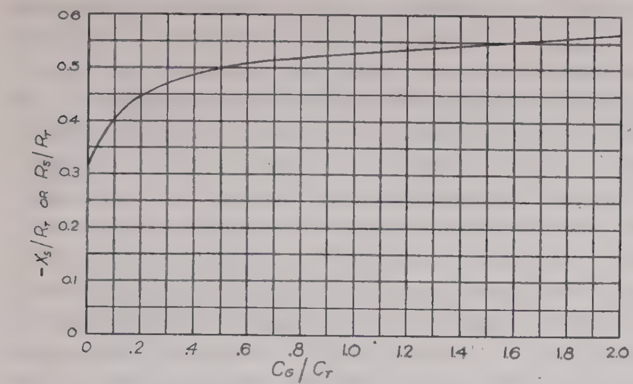

Fig. 10—Equivalent shunt-input reactance and resistance of the resistance-capacitance transmission line.

$-X_s/R_T$ or R_s/R_T versus C_G/C_T. From Fig. 10, for a given value of C_G/C_T, one can then determine the equivalent shunt reactance and resistance seen by the plate of the tube when the resistance-capacitance transmission line is connected.

It should be noted that both the total shunt reactance (X_s in parallel with the reactance of output capacitance plus strays) and the equivalent shunt resistance R_s should be kept high compared to the value of resistance needed to produce the required gain at the frequency specified. If this is not so, i.e.; if the frequency or capacitances are too high, then there will be a loss of gain and, more serious, additional phase shift introduced which will cause the resistance-capacitance transmission line to operate at some other angle than 180 degrees, which will ultimately result in frequency instability and unsteady operation. Thus, the maximum stable frequency obtainable will be limited by whether or not the ratios R_{in}/X_1 and R_{in}/X_s are larger than about 1/10, where R_{in} is the resistance needed to produce the gain required for oscillation (from Fig. 9), X_1 is the parallel combination of X_s and X_0 (the reactance of tube-output capacitance plus strays), R_s and X_s are the values taken from Fig. 10. The maximum frequency is therefore limited by these considerations to a few megacycles, but is nevertheless considerably higher than that obtainable by the lumped-constant oscillator.

C. *Termination in Characteristic Impedance*

An examination of the dip in the curve of Fig. 9 might lead one to wonder what effect will be obtained when the termination is in the characteristic impedance $\sqrt{R/Y}$. This may readily be determined from (8) by setting $I_r\sqrt{R/Y}=E_r$. Expanding the resultant equation for $\sqrt{RY}=\alpha+j\beta$ and setting the j term equal to zero, one obtains the same value of frequency as that given by (17). Substitution of this value into the original equation leads to a ratio E_s/E_r of $-\epsilon^{m\pi}$ (m odd) or -22.3 when $m=1$. Although this is entirely usable, it is in general too high to be of appreciable practical value at the frequencies involved; furthermore the termination conditions will change if the frequency is varied.

D. *Effect of Inductance*

The effect of inductance in the resistance-capacitance transmission line can be easily studied by recalculating the propagation constant $\sqrt{RY}$ using $R(1+jQ)$ instead of R. The resultant equation shows that if Q is less than about 1/3, no effect will be noted. A 50,000-ohm resistor at 1 megacycle would have to have an inductance of 2.65 millihenries to have a Q of 1/3. Even a wire-wound resistor obviously does not have this much inductance, and should therefore perform satisfactorily at frequencies as high as a few megacycles. Experimental results verify this performance.

Conversely, it can also be shown that if Q is greater than about 10, the gain required for oscillation reduces to very nearly unity, and the frequency is reduced to about $1/2Q^2$ of what it was for $Q=0$. This is of no particular consequence for the case under consideration, however.

E. *Design Procedure*

Using the curves given, it is possible to organize the design procedure into 10 steps (in addition to the routine selection of bias and screen voltages, etc.). These steps can be summarized as follows:

1. Select the tube, noting or estimating C_{in}, C_{out}, g_m, and μ.
2. Add to C_{in} and C_{out} appropriate strays to obtain C_G and C_0.
3. Consider the physical limitations of C_T and assign its value; preferably about $5C_G$ for minimum required gain. Find C_G/C_T.
4. Enter the curves at C_G/C_T to find
   (a) Required gain G (Fig. 9)
   (b) fR_TC_T (Fig. 8)
   (c) X_s/R_T (Fig. 10)
   (d) R_s/R_T (Fig. 10)
5. From the desired frequency, C_T, and item 4(b), find R_T.
6. Find X_s and R_s from R_T and items 4(c) and 4(d).
7. Calculate $X_0=1/2\pi fC_0$.
8. Calculate $X_1=X_0/1+(X_0/X_s)$.
9. Calculate $R_{in}=(G/g_m)(1/1-G/\mu)$, using the value of G in 4(a), with appropriate margin.
10. Find

$$R_L = R_{in}/\sqrt{1-(R_{in}/X_1)^2-(R_{in}/R_s)}.$$

This should be reasonably close (within about 5 per cent) to R_{in} for best operation. If it is not, then the chosen frequency is too high for the tube constants involved.

A design based on the above steps was made using a 6AC7 tube, resulting in the circuit of Fig. 11, which has been tested experimentally at a frequency of about 500 kilocycles and found to be satisfactory, giving good wave form. The frequency stability compares favorably with other unstabilized self-excited oscillators operating at the same frequency.

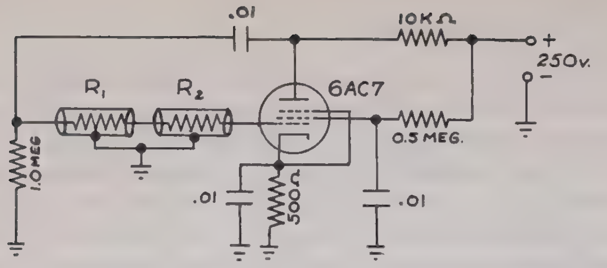

Fig. 11—500-kilocycle phase-shift oscillator. R_1, R_2: two Mallory 50,000-ohm, 10-watt wire-wound resistors covered with metal foil.

F. *Variable-Frequency Operation*

Experimentally it was found that a wire-wound potentiometer with a metal case could be used as the resistance-capacitance transmission line and would give very good results for variable-frequency operation. Other potentiometers were tried also, with foil wrappings for the capacitance element, and found to give equal performance, with frequency variation possible over a very wide range.

A fine control of frequency can also be obtained by connecting a small variable capacitor from grid to ground. This changes the ratio C_G/C_T, thereby changing the frequency by a small amount, as is evident from Fig. 8.

G. *Frequency Stability*

It is evident that, in any oscillator, the frequency-determining components must be as constant as possible if good frequency stability is to be realized. For the phase-shift oscillator using the resistance-capacitance transmission line, the frequency-determining components are the total resistance and total capacitance of the line, and to a certain smaller extent the terminating capacitance of the line. The load and output characteristics of the tube will also affect the frequency unless the input impedance of the line is made very high compared to these parameters.

From an inspection of (19) and Fig. 8, it is clear that in both cases the frequency is inversely proportional to both the total resistance and the total capacitance. A percentage change of either of these latter quantities will therefore result in the same percentage change of frequency, but in the opposite direction. Since the capacitance is only of the order of 50 micromicrofarads, a change of 5 micromicrofarads, for example, would result in a change of 10 per cent in frequency (neglecting the effect on the ratio C_G/C_T). This much change would be quite serious if the frequency were desired constant to, say, 0.01 per cent. Great care must therefore be taken in the construction of the resistance-capacitance transmission line to insure that the temperature coefficient of the resistance is as low as possible, and that the dielectric separating the resistance from its grounded covering be as good as possible. This suggests some sort of a thin-walled polystyrene tubing or similar material, with the resistance wound or sprayed on one side and the capacitance on the other side.

H. *Application to Electrical Measurements*

It has just been mentioned that a percentage change of either the resistance or capacitance of the line results in the same percentage change of frequency, neglecting the change in C_G/G_T. This effect can be utilized to advantage to measure any variable (such as dielectric constant, length, pressure, temperature, humidity, and so on) that affects either the resistance or the capacitance. Frequency changes are easily measured by comparison with a stable frequency standard such as a crystal oscillator, or by other methods.[8,9] In the usual system of this type,[8,9] based on variation of capacitance, the frequency is inversely proportional to the *square root* of the capacitance; while in the phase-shift oscillator the frequency is simply inversely proportional to the capacitance. Thus the phase-shift oscillator will have a sensitivity roughly twice that of inductance-capacitance oscillators, when used for this purpose.

For example, in the measurement or regulation of temperature, a resistance-capacitance transmission line could be constructed, using a material with a high thermal-resistivity coefficient. Large changes in frequency would then result from small changes in temperature.[9]

I. *Other Configurations of the Resistance-Capacitance Transmission Line*

It might be expected that, if the resistance and/or capacitance are tapered in some manner, better operation might result. It has already been shown, for the case of lumped constants, that increasing the impedance toward the grid end of the network reduces the gain necessary for oscillation. This can be shown to be true also for the case of the resistance-capacitance transmission line. By selecting a taper of resistance such that the specific resistance increases as x in Fig. 7 decreases, and a taper of capacitance that decreases at the same time, quite different characteristics can be obtained. In order to investigate these conditions mathematically, however, one must select the equations of R and Y as functions of x, substitute them into (7), and solve the result. This procedure is necessary because R and Y are now functions of x, and since the original solutions of (7) assumed them to be constant, (8) and (9) are no longer valid. This solution may become rather cumbersome, particularly if R and Y are different functions of x, or are not linear functions of x.

By inspection, however, it is evident that we can consider the line as being made up of n resistance-capacitance sections, and if a steep taper is employed, the attenuation will approach $(1/\cos \pi/n)^n$. This obviously approaches unity as n is made very large, indicating a considerable reduction of network loss when the correct taper is employed.

[8] F. E. Terman, "Measurements in Radio Engineering," McGraw-Hill Book Company, New York 18, N. Y., 1935.
[9] Keith Henney, "Electron Tubes in Industry," McGraw-Hill Book Company, New York 18, N. Y., 1937.

J. *Experimental Results*

Sample designs have already been illustrated, using a 6AC7 tube for the high-frequency case and a 6V6 for the low-frequency case. Circuits of these two oscillators are shown in Fig. 11 and Fig. 12, respectively. These circuits have been assembled and tested experimentally.

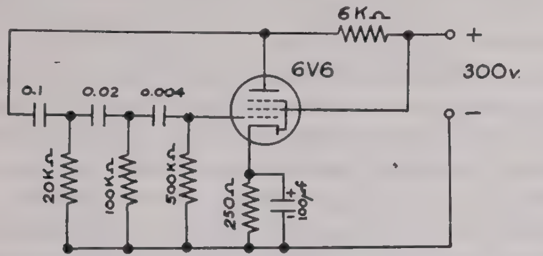

Fig. 12—40-cycle phase-shift oscillator with $K=5$.

The circuit of Fig. 11 oscillated at around 500 kilocycles and gave excellent wave form and stability. Removal of one of the resistors in the phase-shifting element raised the frequency to around 1 megacycle, but the frequency stability was somewhat poorer, which can be attributed to a number of factors. The first and most obvious defect is that the phase-shifting network was rather crude, being comprised of an ordinary wire-wound resistor wrapped with some foil from an electrolytic capacitor. The dielectric separating the resistance from the foil was merely the enamel coating of the resistor, and cannot be expected to have very good insulation properties at 1 megacycle. Another factor is that the wire-wound resistor just mentioned cannot be expected to stay constant with temperature changes, since it is not designed to have zero temperature coefficient. This defect, accompanied probably by other minor factors, caused the oscillator frequency to drift for a time after first being turned on.

The circuit of Fig. 12 oscillated at about 36 cycles. The calculated frequency was nearly 40 cycles. The difference between actual and calculated frequencies can be attributed primarily to the fact that ordinary resistors and capacitors were used, and their accumulated error could easily be 10 per cent. The frequency stability and wave form were excellent.

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Conversion Loss of Diode Mixers Having Image-Frequency Impedance*

E. W. HEROLD†, SENIOR MEMBER, I.R.E., R. R. BUSH†, ASSOCIATE, I.R.E., AND W. R. FERRIS†

***Summary*—The theory of the two-electrode nonlinear mixer for superheterodyne use is developed so as to include the effect of resistive impedance at image frequency. The general theory is applied to the calculation of conversion loss under optimum conditions of matching the intermediate-frequency circuit to the mixer stage. Even when the intermediate-frequency circuit is not matched to the mixer, this matched-impedance conversion loss is important in the determination of the over-all signal-to-noise ratio of a receiver. This loss is computed for an idealized diode with different operating conditions, for various values of image-frequency impedance, and for different values of radio-frequency circuit losses.**

Although the chief effect of different image-frequency impedances is a change in the optimum operating conditions and in the required local-oscillator power, there is also an effect on the minimum conversion loss. Very low impedances or very high impedances result in smallest conversion loss. The impedances often encountered at ultra-high frequency (image-frequency impedance approximately the same as signal-frequency impedance) result in an increase in conversion loss which may be between 0 and 3 decibels, depending on the circuit losses.

I. Introduction

THE PURPOSE of this work is to extend the existing theory of the two-electrode, nonlinear device used as a superheterodyne mixer to include the effects of an essentially resistive impedance at image frequency, and to compute the effect on conversion loss. In many ultra-high-frequency applications, the intermediate frequency is sufficiently low so that the signal input circuit has an appreciable impedance at both the signal frequency and its image. Thus, a complete treatment must include the effect of this additional complication.

The diode and the crystal rectifier are the outstanding examples of the nonlinear devices considered in this paper. Even neglecting impedance at image frequency, their behavior has been more difficult to treat than that of grid-controlled mixers because of the reaction of the intermediate-frequency voltage on the signal circuit. When an impedance is present at image frequency, the treatment is even more complex because, as will be shown, image-frequency currents and voltages are caused both by the modulation of the local-oscillator frequency by the intermediate frequency, and by the modulation of the second harmonic of the local-oscillator frequency by the signal frequency. The image-frequency voltage, in turn, reacts on both signal and intermediate-frequency circuits. However, the complete solution can be reduced to three comparatively simple equations from which numerical results can be

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† R.C.A. Laboratories, Radio Corporation of America, Princeton, N. J.

obtained, and from which an equivalent circuit can be drawn in the form of a 6-pole.

The diode mixer, neglecting image-frequency impedance, has been treated by Strutt[1] and by Strutt and Van der Ziel[2] in incomplete form, and more completely by James and Houldin[3] and by Herold.[4] Peterson and Hussey considered image-frequency impedance in diode or crystal modulator circuits with resistive loads.[5] Two more recent papers discuss reactive loads as well, though not in complete fashion.[6,7] Attention in the present work is confined to resistive loads and, although the equivalent-circuit equations are essentially the same as those of Peterson and Hussey, this paper presents computations of the effect of image-frequency impedance on conversion loss. In view of the importance of the signal-to-noise ratio in ultra-high-frequency receivers, the conversion loss of the diode or crystal mixer is a most important quantity and is of great value in estimating the over-all noise factor of the system.[4]

Let us first explore, in an introductory way, the possible mixing effects in a nonlinear device. When a voltage at signal frequency ω_s is impressed and a local-oscillator voltage of frequency ω_0 is also present, a current at the intermediate frequency, $\omega_{i-f}=(\omega_s-\omega_0)$, is produced. An impedance at this difference frequency leads to an intermediate-frequency voltage. This intermediate-frequency voltage, in turn, modulates currents at oscillator fundamental and causes currents to flow at frequencies $(\omega_0+\omega_{i-f})$ and $(\omega_0-\omega_{i-f})$. The first of these is the same as the signal frequency, but the second is a new frequency which differs from that of the signal by twice the intermediate-frequency and is called the image frequency.[8] The signal-frequency voltage, modulating currents at the oscillator second harmonic, gives rise to a current at frequency $(2\omega_0-\omega_s)$ which again differs from the signal by twice the intermediate-frequency and so is also at image frequency. When the signal circuit is broad, the image-frequency currents give rise to appreciable image-frequency voltages which react on both signal and intermediate-frequency circuits. Thus we must consider, in addition to the local oscillator, currents and voltages at three frequencies: signal, image, and intermediate-frequency, and each of these reacts on the others.

[1] M. J. O. Strutt, "Diode frequency changers," *Wireless Eng.*, vol. 13, pp. 73–80; February, 1936.

[2] M. J. O. Strutt and A. Van der Ziel, "The diode mixer tube, particularly at decimeter wavelengths," *Philips Tech. Rundschau*, vol. 6, pp. 289–299; October, 1941.

[3] E. C. James and J. E. Houldin, "Diode frequency changers," *Wireless Eng.*, vol. 20, pp. 15–27; January, 1943.

[4] E. W. Herold, "Frequency mixing in diodes," Part V of "Some aspects of radio reception at ultra-high-frequencies," by E. W. Herold and L. Malter, Proc. I.R.E., vol. 31, pp. 575–582; October, 1943.

[5] E. Peterson and L. W. Hussey, "Equivalent modulator circuits," *Bell Sys. Tech. Jour.*, vol. 18, pp. 32–48; January, 1939.

[6] H. Meinke, "The behavior of mixing diodes at low and high frequencies," *Elek. Nach. Tech.*, vol. 20, pp. 39–48; February, 1943.

[7] H. F. Mataré, "Input and output resistances of mixing diode," *Elek. Nach. Tech.*, vol. 20, pp. 48–59; February, 1943.

[8] This is identical with the frequency at which a superheterodyne receiver has a second (usually undesired) response. However, in the instance here considered, no image-frequency signals need be impressed to give rise to currents and voltages at that frequency.

In a prior analysis,[4] it was shown that the diode or crystal mixer without image-frequency impedance may be considered as a passive transducer which, at the same time, changes frequency. Thus, when the mixer stage is driven from an antenna, part of the input power is dissipated in the mixer itself while the remainder is available as output power in the intermediate-frequency load. It is clear that, if some of the input power is also converted to image-frequency power, this should represent an additional source of loss. However, when the image-frequency impedance is either zero or infinite its power-absorbing capability is zero; i.e., it can only cause greater conversion loss at intermediate values of impedance, where it approximates a match to the diode. We shall see that these expectations are to some extent fulfilled, but that, in particular instances, a very high image-frequency impedance can actually improve the performance and *decrease* the conversion loss slightly. Unfortunately, in practical cases the net improvement possible is so slight as not to justify the special technique needed.

II. Conversion Theory

The method of analysis to be employed here is similar to one previously used[4] except for the inclusion of the image-frequency impedance. The basic circuit is shown in Fig. 1(a). For convenience in analysis the circuit of

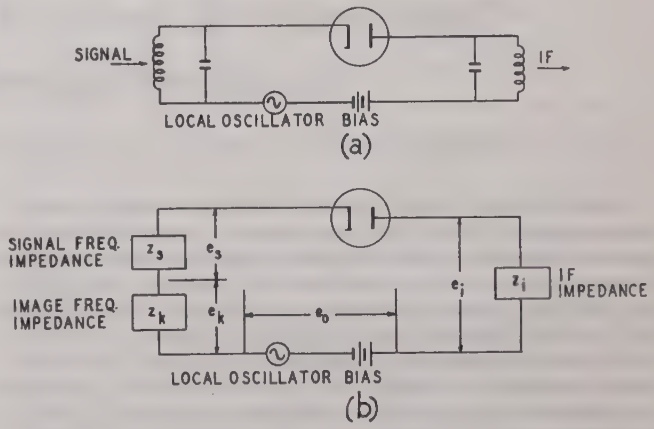

Fig. 1—Diode-mixer circuit and equivalent circuit showing impedances to signal, image, and intermediate frequencies.

Fig. 1(b) has been drawn. In this equivalent circuit, Z_s represents the total impedance at signal frequency of all the circuits connected across the diode (the total impedance presumably derives mainly from the signal input circuit). Similarly Z_k represents the total impedance at image frequency (which may also derive from the signal input circuit) and Z_i the total intermediate-frequency impedance. The impedance at local-oscillator frequency need not be included so long as a definite bias and local-oscillator voltage exists (shown as e_0 in the figure).

The voltages at signal, image, and intermediate frequencies are shown as e_s, e_k, and e_i respectively. The current in the diode or crystal is

$$i = f(e) = f(e_0 + e_s + e_k + e_i) \tag{1}$$

where $f(e)$ is the characteristic (assumed to be single valued) of the nonlinear device. When the signal-frequency voltage and the resulting image and intermediate-frequency voltages are small, a Taylor's expansion about e_0, the instantaneous voltage (which may have any value), gives

$$i = f(e_0) + (e_s + e_k + e_i)f'(e_0) + \cdots. \tag{2}$$

But since $f'(e_0) = \partial i/\partial e\,|_{e=e_0} = g(e_0)$

it is seen that

$$i = f(e_0) + (e_s + e_k + e_i)g(e_0) + \cdots \tag{3}$$

where $g(e_0)$ represents the instantaneous conductance of the nonlinear device. Of the resulting current, it is clear that the first term represents only direct current, and currents at local-oscillator frequency and at harmonics of this frequency. In fact, it represents just the currents which would flow if no signal-frequency, image-frequency, or intermediate-frequency voltages were present. The second term is the one of major interest, since it contains the other currents of interest.[9] We then write

$$i = (e_s + e_k + e_i)g \tag{4}$$

where it is understood that g is a function of e_0; i.e., the instantaneous conductance g varies periodically in time at the local-oscillator frequency. If it is understood that the local-oscillator voltage is a cosine function, we may write the following Fourier series for g:

$$g = g_0 + 2\sum_{n=1}^{\infty} g_{cn} \cos n\omega_0 t \tag{5}$$

where

$$g_0 = \frac{1}{2\pi}\int_0^{2\pi} g\,d(\omega_0 t) \tag{6}$$

$$g_{cn} = \frac{1}{2\pi}\int_0^{2\pi} g \cos n\omega_0 t\,d(\omega_0 t). \tag{7}$$

It is now assumed that the signal-frequency, image-frequency, and intermediate-frequency voltages are sine waves, and that the impedances, across which these voltages appear, are pure resistances. Thus, we write

$$\begin{aligned} e_s &= E_s \sin \omega_s t \\ e_i &= E_i \sin (\omega_i t + \phi_i) \\ e_k &= E_k \sin (\omega_k t + \phi_k) \end{aligned} \tag{8}$$

$$\begin{aligned} i_s &= I_s \sin \omega_s t \\ i_i &= -\,I_i \sin (\omega_i t + \phi_i) \\ i_k &= -\,I_k \sin (\omega_k t + \phi_k) \end{aligned} \tag{9}$$

where ϕ_i and ϕ_k are arbitrary phase angles and the two negative signs allow for the fact that e_i and e_k are voltage drops. By an extension of the procedure previously used,[4] it may be shown that the phase angles are both equal to π.

In the general case, the desired intermediate-frequency may be obtained by conversion at any local-oscillator harmonic, so that $\omega_i = \pm(n\omega_0 - \omega_s)$. In the usual case $n=1$ but the analysis is just as simply carried out for the general case. Thus $\omega_s = n\omega_0 \pm \omega_i$; $\omega_k = n\omega_0 \mp \omega_i$. From these, it is evident by eliminating ω_i that $\omega_s = 2n\omega_0 - \omega_k$; $\omega_k = 2n\omega_0 - \omega_s$. Equations (5) and (8) may now be inserted in (4), the trigonometric products expanded in sum and difference angles, and use made of (9), giving

$$\begin{aligned} I_s &= -\,g_{cn}E_i + g_0E_s + g_{c2n}E_k \\ 0 &= +\,g_{cn}E_i - g_{c2n}E_s - (g_0 + g_k)E_k \\ 0 &= -\,(g_0 + g_i)E_i + g_{cn}E_s + g_{cn}E_k. \end{aligned} \tag{10}$$

These three equations were obtained by separating the current in (4) into its components and neglecting all frequency terms, except those in ω_s, ω_k, and ω_i, since it is assumed that impedances to higher frequencies are negligible. g_k and g_i are the image-frequency and intermediate-frequency circuit conductances, respectively, defined by $g_k = I_k/E_k$; $g_i = I_i/E_i$. It will be observed that three coefficients, g_0, g_{cn} and g_{c2n}, of the Fourier series for the mixer conductance (5) enter into the results. These coefficients may be evaluated from (6) and (7) or by other methods.[10] Examination of the equations shows that the mixer may be represented by the equivalent circuit, shown in Fig. 2 and first given by Peterson and Hussey.[5]

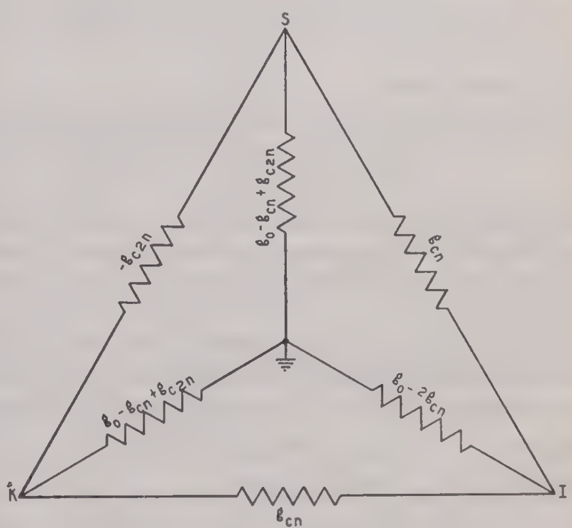

Fig. 2—Six-pole equivalent circuit of diode mixer with image-frequency impedance. Signal is applied between S and ground, the intermediate-frequency impedance is connected between I and ground, and the image-frequency impedance between K and ground.

III. Matching Conductances and Conversion Loss

The equivalent circuit did not prove to be useful in analyzing the conversion characteristics of the mixer. A direct solution of (10) will give for the output-to-input voltage ratio

$$\frac{E_i}{E_s} = \frac{g_{cn}(g_0 + g_k - g_{c2n})}{(g_0 + g_k)(g_0 + g_i) - g_{cn}^2}. \tag{11}$$

Defining the input conductance, g_{in}, as I_s/E_s, we also obtain from (10)

[9] The assumption under which higher order terms in the Taylor's expansion can be neglected is that the signal-frequency, image-frequency, and intermediate-frequency voltages are sufficiently small to justify dropping terms which would result from expanding the derivatives of the conductance into their respective Fourier series.

[10] A simple analysis from the characteristic curve is given in the paper, E. W. Herold, "The operation of frequency converters and mixers," Proc. I.R.E., vol. 30, pp. 84–103; February, 1942.

$$g_{in}=\frac{g_{cn}^2(2g_{c2n}-2g_0-g_k)-(g_0+g_i)(g_{c2n}^2-g_0^2-g_0g_k)}{(g_0+g_k)(g_0+g_i)-g_{cn}^2}. \quad (12)$$

The ratio of power output to input is

$$(P_0/P_{in}) = (E_i^2g_i/E_s^2g_{in}). \quad (13)$$

Maximization of (13), with respect to g_i, gives

$$(P_0/P_{in})_{max} = (\sqrt{\eta} - \sqrt{\eta - 1})^2 \quad (14)$$

where

$$\eta = \frac{[1 + g_k/g_0 - (g_{cn}/g_0)^2][1 + g_k/g_0 - (g_{c2n}/g_0)^2]}{(g_{cn}/g_0)^2(1 + g_k/g_0 - g_{c2n}/g_0)^2}. \quad (15)$$

This maximum power ratio occurs when

$$g_i^2 = \left(g_0 - \frac{g_{cn}^2}{g_0 + g_k}\right)\left[\frac{g_{cn}^2(2g_{c2n} - 2g_0 - g_k)}{g_0^2 + g_0g_k - g_{c2n}^2} + g_0\right]. \quad (16)$$

In the limit, when g_k becomes infinite (zero image-frequency impedance) (14) and (16) reduce to (17) and (19) in Herold's paper.[4]

Moreover, the maximum power ratio in (14), as a function of the image-frequency conductance g_k has a minimum value of

$$\left[\left(\frac{P_0}{P_{in}}\right)_{max}\right]_{min} = \left(\frac{g_0}{g_{cn}}\right)^2 \left\{\frac{(g_{c2n}/g_0)[(g_{cn}/g_0)^2 - (g_{c2n}/g_0)]}{1 - (g_{c2n}/g_0)}\right\} \quad (17)$$

which occurs when

$$\frac{g_k}{g_0}=\frac{(g_{c2n}/g_0)[(g_{cn}/g_0)^2+(g_{c2n}/g_0)^2]-2(g_{cn}/g_0)^2(g_{c2n}/g_0)^2}{2(g_{c2n}/g_0)-[(g_{cn}/g_0)^2+(g_{c2n}/g_0)^2]^2}-1. \quad (18)$$

This shows that, if the image-frequency conductance is of a suitable value, it absorbs sufficient power to reduce the normal power transfer to a minimum value as given by (17).

The analysis thus far has assumed that the input circuit has no losses. If we let the input circuit have certain losses represented by a shunt conductance, g_s, the power ratio is

$$P_0/P_{in} = E_i^2g_i/E_s^2(g_{in} + g_s) \quad (19)$$

which has a maximum value of

$$(P_0/P_{in})_{max} = (\sqrt{\eta'} - \sqrt{\eta' - 1})^2 \quad (20)$$

where

$$\eta' = \eta\left[\frac{(1 + (g_k/g_0)(1 + (g_s/g_0) - (g_{c2n}/g_0)^2}{1 + (g_k/g_0) - (g_{c2n}/g_0)}\right]. \quad (21)$$

The function in (20) also has a minimum as g_k is varied between zero and infinity.

A case of particular interest is that for which the total image-frequency conductance is equal to the total signal-frequency conductance. This occurs in a mixer having a broad input circuit and a low intermediate frequency. In the previous analysis it was assumed that the signal source, which may be thought of as consisting of a current generator of current i_a, and a shunt conductance g_a, was matched to the mixer input; viz.,

$$g_a = g_{in} + g_s. \quad (22)$$

The power input to the mixer P_{in} was then equal to the "maximum available power," from the source, given by

$$P_a = i_a^2/4g_a. \quad (23)$$

By this assumption, it was necessary to maximize (19) with respect to g_i only, since it was already maximized with respect to the source conductance g_a by the above reasoning.

However, if we make $g_k = g_a + g_s$, we must consider the source conductance when determining the minimum conversion loss.[11] In all cases, the conversion loss will be determined by the ratio of power output to the "maximum available power" from the source. The voltage across the input will be

$$E_s = i_a/(g_a + g_s + g_{in}). \quad (24)$$

The output power is

$$P_0 = E_i^2g_i \equiv E_i^2g_i(E_s^2/E_s^2) = \frac{E_i^2}{E_s^2}g_i\frac{i_a^2}{(g_a + g_s + g_{in})^2} = \frac{i_a^2}{4g_a}\frac{4g_ag_i}{(g_a + g_s + g_{in})^2}\frac{E_i^2}{E_s^2}. \quad (25)$$

The conversion loss will then be given by

$$P_0/P_a = 4g_ag_i/(g_a + g_s + g_{in})^2E_i^2/E_s^2. \quad (26)$$

If g_k is replaced by (g_a+g_s) in (11) and (12) and the results inserted in (26) and the expressions simplified, we will have

$$\frac{P_0}{P_a} = 4g_ag_i\left[\frac{g_{cn}}{(g_0 + g_i)(g_{c2n} + g_0 + g_a + g_s) - 2g_{cn}^2}\right]^2. \quad (27)$$

This expression is readily maximized with respect to both g_a and g_i, giving

$$(P_0/P_a)_{max} = 1/2(\sqrt{\eta''} - \sqrt{\eta'' - 1})^2 \quad (28)$$

where $\eta'' = (1 + g_{c2n}/g_0 + g_s/g_0)/2(g_{cn}/g_0)^2. \quad (29)$

The greatest real value that the function in (28) can have is 1/2, corresponding to $\eta''=1$, while the greatest real value of the analogous expression in (14) and (20) is 1. Thus, if the total image-frequency and signal-frequency conductances are equal, the minimum conversion loss is 3 decibels, whereas in the other cases the conversion loss may be as low as 0 decibels.

IV. Idealized Diode

From the above analysis, we can now determine the proper matching impedance and the conversion gain to be expected, once we know the conductance coefficients of the diode, the circuit losses, and the image-frequency impedance. The conductance coefficients are, to a great extent, within our control, since they depend not only upon the tube characteristic, but also upon the bias and the amplitude of the local-oscillator voltage. We have little control over the circuit losses, but usually try to keep them as low as possible. The image-frequency impedance is mainly governed by the selectivity of the input circuit and the choice of the intermediate frequency. So far, it is not at all obvious that the image-frequency impedance should be made as low as possible.

[11] This was pointed out to us by D. O. North, who also suggested the method of analysis which follows.

Since the implications of our analysis are not clear-cut, at least a qualitative picture will be obtained by considering the simple idealized diode examined by Herold,[4] who neglected image-frequency impedance. The plate

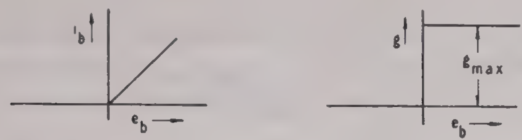

Fig. 3—Plate-current and conductance characteristics of idealized diode.

current and conductance characteristics of such a diode are shown in Fig. 3. The coefficients are

$$g_0 = g_{max}\epsilon$$
$$g_{cn} = g_{max} \sin n\pi\epsilon/n\pi$$
$$g_{c2n} = g_{max} \sin 2n\pi\epsilon/2n\pi$$

where ϵ is the fraction of the cycle during which the diode is conducting. We will consider only the case of conversion at oscillator fundamental; i.e., when $n=1$. We may now calculate the power ratio in (14) and (20) as a function of the ratio of image-frequency conductance to maximum diode conductance g_k/g_{max} for different values of ϵ, or of the ratio of bias to peak oscillator voltage E_b/E_0, since $\pi\epsilon=\cos^{-1}(E_b/E_0)$. This has been done for $\pi\epsilon$ equal to $\pi/4$, and 0.3, for zero circuit losses and for losses such that g_s is 1 per cent and 5 per cent of g_{max}. These curves are shown in Fig. 4. The minima of conversion gain, described in Section III, are clearly shown as maxima of conversion loss in the figure.

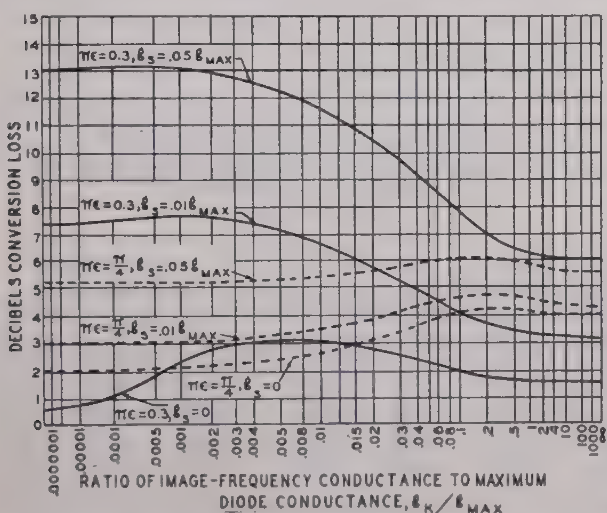

Fig. 4—Mixer conversion loss of idealized diode as a function of image-frequency conductance.

It is, perhaps, more informative to calculate the conversion loss as a function of E_b/E_0 for different values of image-frequency conductance and circuit losses. This has been done for four values of image-frequency conductance: (a) infinite image-frequency conductance or zero impedance, the case considered previously by Herold; (b) total image-frequency conductance equal to total signal-frequency conductance, the case of a broad input circuit and a low intermediate frequency, as expressed by (28) and (29); (c) image-frequency conductance equal to input circuit loss conductance g_s, as might occur if g_k were reduced to a low value by introducing a circuit (tuned to the image frequency) having losses roughly equal to the input circuit losses; and (d) zero image-frequency conductance or infinite impedance, the limiting case of (c). Three values of input-circuit loss conductance have been used: (1) $g_s=0$; (2) $g_s=0.01\ g_{max}$; and (3) $g_s=0.05\ g_{max}$. Figs. 5, 6, and 7

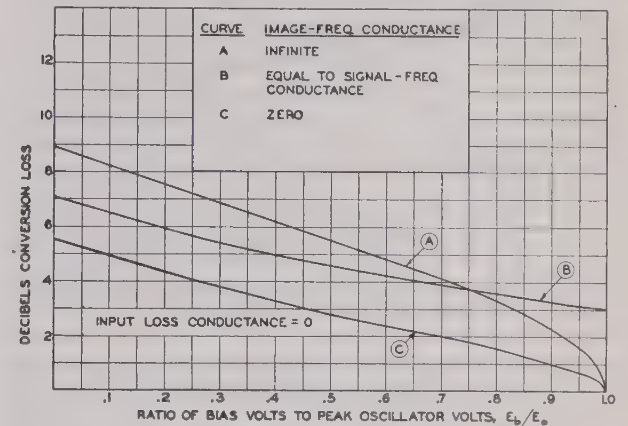

Fig. 5—Mixer conversion loss of idealized diode for no losses in input circuit.

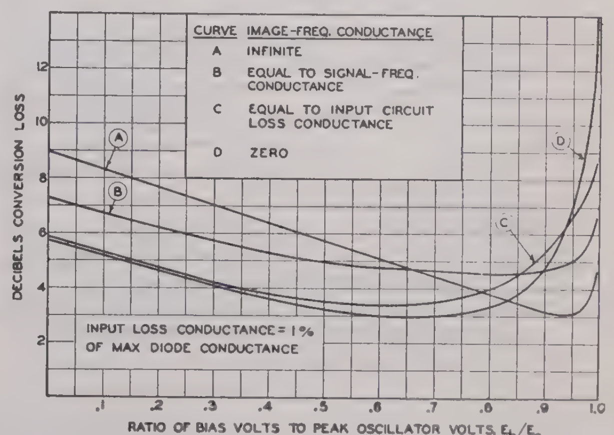

Fig. 6—Mixer conversion loss of idealized diode for input-circuit-loss conductance equal to 1 per cent of maximum diode conductance.

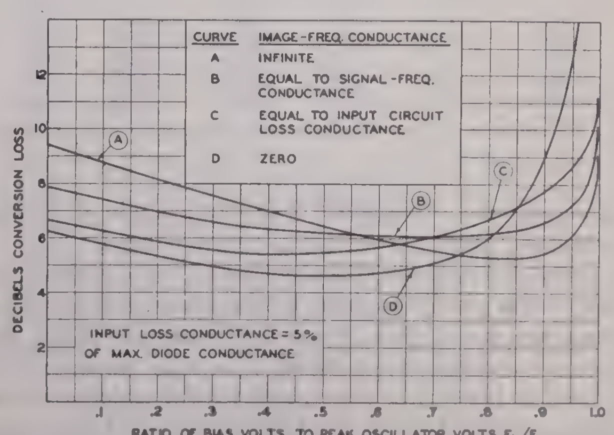

Fig. 7—Mixer conversion loss of idealized diode for input-circuit-loss conductance equal to 5 per cent of maximum diode conductance.

are for zero, 1 per cent, and 5 per cent losses, respectively. These curves have been replotted, to aid comparison, in Figs. 8, 9, 10, and 11 for the four above values of image-frequency conductance. The data used in plotting these curves are given in Table I.

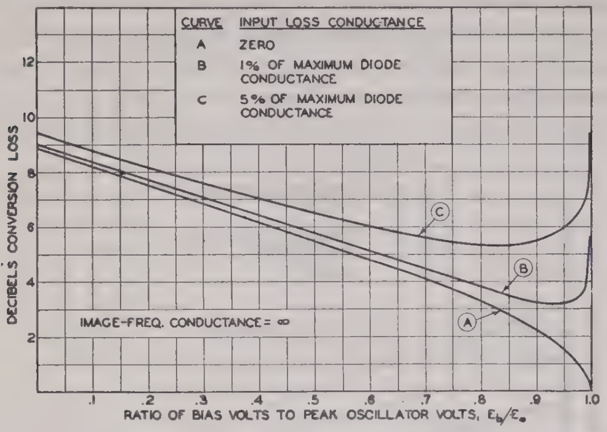

Fig. 8—Mixer conversion loss of idealized diode for infinite image-frequency conductance.

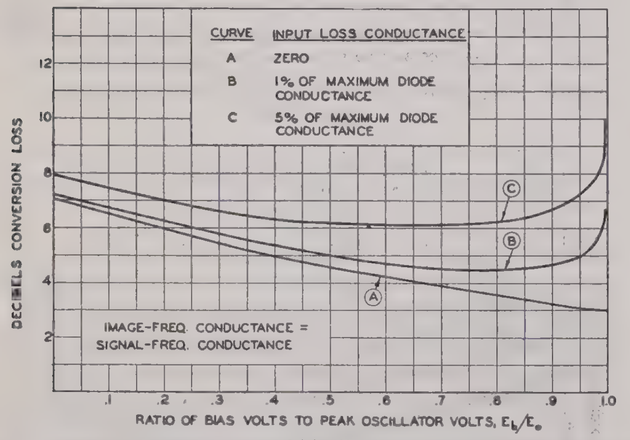

Fig. 9—Mixer conversion loss of idealized diode for image-frequency conductance equal to signal-frequency conductance.

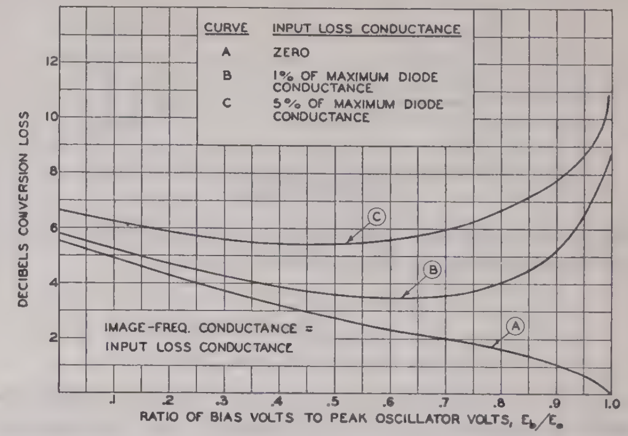

Fig. 10—Mixer conversion loss of idealized diode for image-frequency conductance equal to input-circuit-loss conductance.

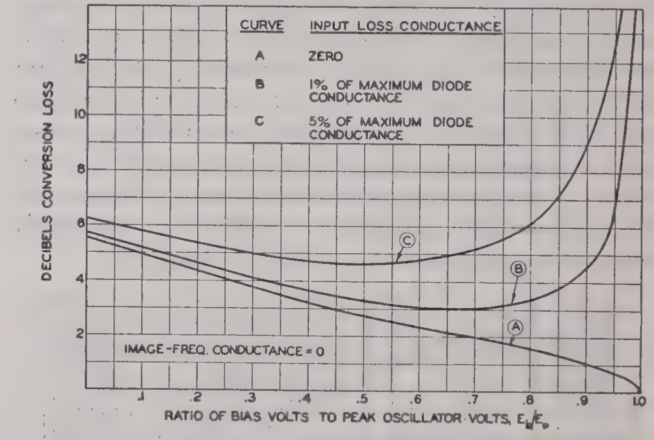

Fig. 11—Mixer conversion loss of idealized diode for zero image-frequency conductance.

TABLE I

Decibels Mixer Conversion Loss of Idealized Diode

E_b/E_o	$g_k \to \infty$ $g_s \to 0$	∞ 1 per cent	∞ 5 per cent	g_a+g_s 0	g_a+g_s 1 per cent	g_a+g_s 5 per cent	g_s 0	g_s 1 per cent	g_s 5 per cent	0 0	0 1 per cent	0 5 per cent
0	8.89	9.00	9.43	7.08	7.25	7.93	5.54	5.80	6.67	5.54	5.71	6.26
0.300	6.82	7.01	7.59	5.45	5.73	6.65	3.77	4.25	5.61	3.77	4.09	4.98
0.500	5.53	5.75	6.53	4.59	5.01	6.19	2.80	3.61	5.45	2.80	3.31	4.64
0.707	4.05	4.43	5.61	3.91	4.65	6.11	1.96	3.55	6.02	1.96	2.98	5.25
0.866	2.66	3.41	5.32	3.38	4.61	6.45	1.22	4.65	7.37	1.22	3.91	7.62
0.955	1.51	3.16	6.06	3.12	5.01	7.33	0.67	6.73	8.76	0.67	7.43	13.2
0.995	0.50	4.67	9.00	3.01	6.36	9.95	0.18	8.48	10.75	0.18	19.8	26.6
1.000	0	∞	∞	3.01	∞	∞	0	∞	∞	0	∞	∞

From these curves, it is observed that the best operating point is greatly changed by the presence of an image-frequency impedance. With zero image-frequency impedance (infinite conductance), the case considered by Herold in his earlier paper, and with reasonable circuit losses, we would expect the least conversion loss to occur when the diode is conducting during a small fraction of the cycle. However, the presence of image-frequency impedance causes minimum conversion loss to occur when the tube is conducting in the neighborhood of one third or one quarter of a cycle. This implies appreciably greater power consumption from the local oscillator.

Moreover, it is easily seen that either zero or infinite image-frequency impedance will give, at the optimum operating angle, lower conversion loss than will an image-frequency impedance equal to the signal-frequency impedance. The difference is very little if the circuit losses are high, but approaches 3 decibels, as shown in Fig. 5, if the losses are very low.

Practically nothing is to be gained by building up the image-frequency impedance to a very high value, as

compared to zero image-frequency impedance. In addition to the greater local-oscillator power required, the presence of high image-frequency impedances may be undesirable from the standpoint of response to unwanted incoming signals. Thus, the design trend of lowering the image-frequency impedance of diode and crystal mixers by the use of a high intermediate frequency seems justified.

V. Practical Diodes

Under any specific set of conditions, the conversion loss may be calculated from (20) after the conductance coefficients have been determined. James and Houldin[5] have determined these coefficients for 3/2 power and exponential characteristics. Herold[10] has described a "7-point formula" which may be applied to empirical characteristics. However, the conclusions drawn from the examination of the idealized diode are qualitatively applicable to practical diodes.

VI. Idealized Crystal Characteristic

Crystal mixers are basically very similar to diode mixers. From the standpoint of conversion loss, the crystal differs from the diode only because the conductance in the negative direction, although small, is not zero. It is helpful to think of the crystal as a diode shunted by a finite resistance. This picture is strictly correct for an idealized characteristic such as is shown in Fig. 12.

Fig. 12—Plate-current and conductance characteristics of idealized crystal.

It seems fairly clear that no improvement in performance could be expected by the introduction of conductance in the negative direction. This is certainly true in the absence of image-frequency impedance. A more detailed analysis shows that this is also true when image-frequency impedance is present, provided operating conditions are optimized. Thus, the only important conclusion is that which would be expected, namely, conductance in the negative direction modifies the optimum operating point for the mixer, increases the conversion loss, and consumes local-oscillator power.

Electrical Testing of Coaxial Radio-Frequency Cable Connectors*

CHANDLER STEWART, JR.†, ASSOCIATE, I.R.E.

Summary—**Tests for effective characteristic impedance, reflection coefficient, and power factor of coaxial radio-frequency cable connectors are described, and mathematical derivations of formulas given. Results of these tests on a few typical connectors are tabulated.**

Introduction

MILITARY requirements have recently brought about a tremendous increase in quantity and types of radio and electronic equipment operating in the very-high-frequency and ultra-high-frequency ranges. Because of the greater effect of coaxial radio-frequency cable-connector characteristics upon the efficiency and performance of equipment at these frequencies than at lower frequencies, the design and testing of these connectors naturally become of increased importance. In fact, instances have occurred recently where an equipment would perform satisfactorily with a certain type of antenna cable connector, but would not function properly with connectors of the same type but of different manufacture. Consequently, it has been necessary to set up methods for evaluating the relative merit of coaxial radio-frequency cable connectors on the basis of their electrical characteristics. Some of these methods, which have been employed at the Radio and Radar Subdivision, Air Technical Service Command, Wright Field, Ohio, will be described.

Of these, the two characteristic impedance tests were developed by the author. The other tests are adaptations of methods previously used. The mathematical analysis of the behavior of errors in the insertion method was carried out by Lieutenant Saul Gorn. Study of the problems of connector testing was also participated in by Mr. Daniel Robb and Miss Carolyn Parker.

The characteristic impedance of a connector aids the designer and manufacturer in modifying the connector to improve the impedance match to the cable with which it is used. Since this is an effective value, based upon the assumption that the connector has uniform characteristics along its length, there may be a variation of test results with frequency, which, however, has been found to be small in the samples so far tested. The two methods for determining characteristic impedance to be described are the "insertion" method and the "null-shift" method.

The reflection coefficient of a connector is a measure of the effect of the impedance mismatch between the connector and a given cable at a given frequency, but it

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† Headquarters, Army Air Forces, Air Technical Service Command, Wright Field, Dayton, Ohio.

gives no clues for correcting this mismatch. In order to transmit power efficiently over a coaxial cable, reflections at the connectors must be kept to a minimum, so that the connectors should be designed and inspected for a low reflection coefficient. Since reflection coefficient varies greatly with frequency and cable impedance, these must be specified for the reflection coefficient tests. A frequency of 500 megacycles and an impedance of 50 ohms are commonly used for this purpose. Reflection coefficient is obtained by the "insertion method," mentioned above in connection with characteristic impedance tests. The two quantities can be obtained simultaneously by this method, as will later be shown.

The power factor of the connector dielectric gives a relative indication, for a given type of connector, of the power absorbed (instead of reflected) in the connector, and aids in the detection of flaws in the dielectric, etc. This test is most conveniently made at frequencies from 1 to 100 megacycles, and the results may vary somewhat with frequency.

I. Insertion Method of Reflection and Characteristic Impedance Test for Coaxial Connectors

A. *Summary (See Fig. 1)*

For this test, the arrangement of apparatus is as shown in Fig. 1. The test specimen is inserted between the line and lossy cable as shown, and the indicator (which is essentially a radio-frequency detector) is slid

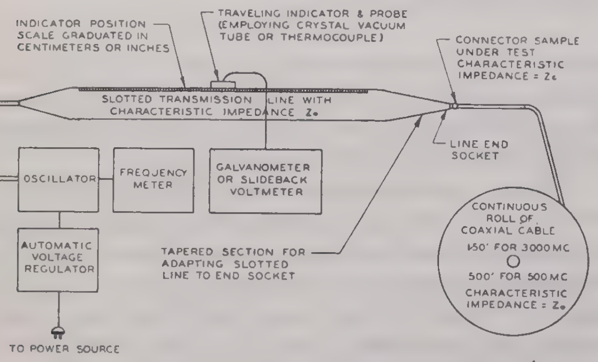


Fig. 1—A typical arrangement of apparatus for insertion test.

along the line, and the maximum and minimum galvanometer indications observed, as well as the position of the indicator for minimum indication. These readings, corrected from a calibration chart, together with the indicator position for minimum galvanometer deflection with the line short-circuited, are essentially applied to the curves of Fig. 2 to obtain the reflection coefficient and effective characteristic impedance of the specimen.

B. *Equipment required*

(1) A slotted transmission line whose end socket has been tailored to have the same characteristic impedance as the line, as will be described.

(2) A standing-wave indicator for the slotted line.

(3) A radio-frequency source with regulated power supply. Adjustment of power should be made to a value found in practice to be most suitable. About 1 watt was found adequate in most cases.

(4) Frequency meter.

(5) A continuous roll of flexible coaxial cable, long enough to absorb reflections, and of the same type with which the connector will be used in its most critical application. Its characteristic impedance must, of course, be the same as that of the slotted line. Lossy line is in general unsatisfactory, since it usually has conductor diameters differing from those of the cables for which the connectors were designed, so that the reflection at the connector would be affected. In practice, 500 feet of ordinary polyethylene dielectric cable will be found sufficient for 500-megacycle tests, and 150 feet long enough for tests at 3000 megacycles.

(6) A silver-plated short-circuiting plug for the slotted-line end socket.

C. *Calibration of the Indicator*

For the condition of a short-circuited slotted line, we substitute $R=0$ and $X=0$ in (2) of Section IV, which follows, yielding

$$E_s = I_R Z_0 \sin \beta l.$$

Since β and I_R are constant for a given set of conditions, E_s would vary sinusoidally along the line. This fact is used as a basis for calibrating the indicator at a frequency near that of the measurement. Sin βl is plotted as a function of the potentiometer setting, and this curve is used in determining relative values of E_s from the potentiometer settings.

D. *Tailoring of the Slotted-Line End Socket*

It is obvious that in all work involving a slotted line, the characteristic impedance of the line must be uniform over the region including the points at which voltage indications are obtained and the terminating load. This means that steps must be taken to insure that the socket at the load end of the line has the same characteristic impedance as the line itself. The end socket used is similar to a type SO-239 (Navy type—49194), with the exception that its characteristic impedance has been raised to 50 ohms, the characteristic impedance of the slotted line. Its end capacitance should be approximately the same as that of the male end of the fitting since the end is of similar construction. It is shown in (66) of Section IV that short-circuiting the end socket should shift the line null position by ($\lambda/4-1.5\ C_{\mu\mu f}$) centimeters. If the actual shift is less than this, the socket impedance is too small, and can be improved by drilling holes in or cutting away part of the dielectric, reaming out the outer conductor, or machining down the inner conductor. If the null shift is more than ($\lambda/4-1.5\ C_{\mu\mu f}$) centimeters, the socket impedance is too large and can be lowered by adding dielectric, or reducing the ratio of conductor diameters.

E. *Checking for Discrepancy Between Slotted-Line and Cable Characteristic Impedance*

One method of accomplishing this, which was tried with satisfactory results, is as follows: (1) Ream out the dielectric of a coaxial plug, type PL-259. (2) Run the cable end completely through the plug shell, soldering the outer braid end to the large end of the plug. (3) Remove insulation from the plug pin, and solder it to the inner-cable strands, so that it is supported only by them. (4) If this is properly done, the plug can now be connected to the line so that the cable end is brought directly in contact with the end of the line, reducing reflections to a minimum. (5) The standing-wave ratio is then checked. When cable and line were well matched, a ratio of at least 0.98 was readily obtained with the equipment used.

F. *Determining the Short-Circuit Null Position*

Before the equipment can be used for connector tests, the position of the indicator for minimum voltage must be determined for the condition of the short-circuiting plug in place at the end of the line. This reading is taken at the test frequency, while the exact frequency is found on the frequency meter. The radio-frequency source should be adjusted to this frequency in subsequent work where this null position is used.

G. *Choice of Operating Frequency for Characteristic Impedance Tests*

The errors inherent in the equipment will be greatly magnified unless a test frequency near an optimum value is used. Small variations in line constants or in the coupling between the indicator probe and the inner conductor of the slotted line will obviously result in an error in the standing-wave ratio r. Also, they may cause an error in ΔS by virtue of an effective shift in the position of the voltage minimum. Limits of the error in determining Z_c due to these effects are represented by the first and second terms, respectively, of the following formula from Section IV:

$$\zeta = \left| Z_0^2/Z_c^2 - Z_c^2/Z_0^2 \right| \cdot \left| r/(1-r^2) \right| \eta + 2\left| Z_0^2/Z_c^2 - 1 \right| \frac{r\sqrt{(r - Z_c^2/Z_0^2)\left[1 - r(Z_c^2/Z_0^2)\right]}}{(1-r)\sqrt{1-r^4}} \sqrt{\eta}. \quad (42a)$$

It will be shown later that ζ is smaller the nearer $\beta_c l_c$ approaches $(n+\frac{1}{2})\pi$. Also, it will be shown there that the nearer Δs approaches 0 or ± 0.25, the lower the resulting error.

In actual practice, operation at a frequency far removed from that of best accuracy may lead to practically useless results. For example, for the condition of $Z_0=50$, $Z_c=25$, and $\eta=0.02$, from Fig. 2 and (42a) we obtain the following maximum errors: When $\Delta S=0.037\lambda$, $\zeta=0.041$, which is permissible for this work. But when $\Delta S=0.118\lambda$, $\zeta=8.1$, which, of course, renders the test results worthless.

Equation (47a) shows the relation between a frequency of best accuracy, f_0, (corresponding to $\beta_c l_c=\pi/2$) an arbitrarily chosen trial frequency f_1, which should be less than f_0, and the r and ΔS obtained at f_1 on a given connector. This equation has been graphed by plotting as dashed lines in Fig. 2. A rough estimate of f_0 can be obtained from (45), Section IV: $f_0=7.5\cdot 10^9/l_c\sqrt{\epsilon}$.

If the trial frequency is chosen to be far enough below the value of f_0 estimated from the above formula to insure its being less than the true f_0, the dashed lines in Fig. 2 can be used as a basis for changing the test frequency to one of more practical value. However, if the trial frequency is too far below the best frequency, the determination of the latter will be inaccurate and more than one successive approximation will be required.

H. *Taking Data on Individual Connector Samples*

After the equipment has been prepared as just explained, the connector to be tested is soldered to the cable end in the same fashion as it is expected to be used in service. The connector is then attached to the slotted line, and the standing-wave ratio and the position of the point of minimum voltage are noted. For testing adaptors, the adaptor is inserted between the slotted line and a plug which is made up as just described under Section I, E, above. The effective characteristic impedance of the connector under test is read on a graph of (24a) from Section IV.

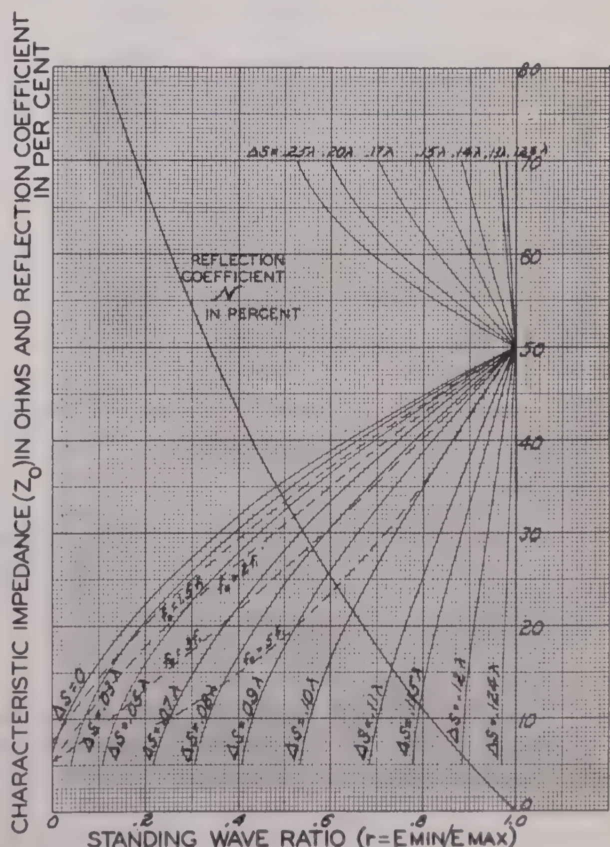


Fig. 2—Characteristic impedance and reflection coefficient in terms of test data.

$$Z_c = Z_0\sqrt{[r - \tan^2(2\pi\Delta s/\lambda)]/[1 - r\tan^2(2\pi\Delta S/\lambda)]}. \quad (24a)$$

The solid curves in Fig. 2 are from this equation, for the condition of $Z_0 = 50$ ohms.

I. *Checks on the Data Obtained*

For test frequencies less than f_0, as shown in (44), $\Delta S/\lambda$ must lie between zero and 0.125 or between -0.25 and -0.125. Checking each data point for this requirement affords a partial check on the precision of results.

Also, from Section IV E(1), the data points must lie outside the region in Fig. 2 bounded by the curves $\Delta S = 0$ and $\Delta S = 0.25\lambda$. This gives another partial check on the results.

It should be understood that the results obtained by this method represent the effective characteristic impedance of the entire junction, and are influenced by the manner in which cable is connected to the plug or socket. Evidence of the extent of the dependence of the electrical characteristics of the junction upon the type of soldering job is shown in Table I, where reflection coefficients obtained on the same connectors soldered by different individuals are compared.

TABLE I

Reflection coefficients of a few type SO-239 sockets at 500 megacycles. Measurements were made by two technicians after each soldering of the socket to the long cable.

Sample Number	Manufacturer	Connection soldered by				
		Stewart	Mutrux	Selke	Dunlevy	Feher
44A	L	0.110 0.109	0.136 0.147	0.138 0.143	0.148 0.147	
44B	L	0.135 0.133	0.140 0.138	0.111 0.111	0.128 0.133	
44C	L	0.136 0.136	0.114 0.115		0.114 0.114	
55A	S	0.112 0.115	0.149 0.149	0.121 0.121	0.124 0.131	0.096 0.092
55B	S	0.164 0.161	0.097 0.095	0.122 0.123	.107 .107	
55C	S	0.085 0.089	0.120 0.120	0.212 0.212	0.125 0.125	
56A	A	0.157 0.157	0.146 0.149			
56B	A	0.153 0.152	0.134 0.136	0.155 0.150	0.125 0.121	
56C	A	0.166 0.166	0.157 0.157		0.134 0.125	

J. *Reflection-Coefficient Test*

The standing-wave ratio obtained on the line is used to determine the reflection coefficient of the plug-and-socket combination, by the formula[1] $N = (1-r)/(1+r)$. This equation has also been plotted in Fig. 2.

K. *Evaluation of the Insertion Method*

In Table II are listed experimental data, and, for comparison, estimated values obtained from the following formula, which is in general use:

$$Z_c = (138 \log_{10} c/b)/\sqrt{\epsilon} \text{ ohms}$$

where b = diameter of inner pin

c = inner diameter of outer shell in same units.

TABLE II

Specimen		Test Frequency	Characteristic Impedance		Reflection Coefficient N	Power Factor
Type	Manufacturer		Ohms	Method		
M-359	A	300	37.1	null-shift		
		500	35.0	insertion	0.25	
		800	36.1	insertion		
M-359A	A	300	43.1	null-shift		
		500	43.8	insertion	0.14	
		800	44.3	insertion		
M-359	B	300	44.1	null-shift		
		500	43.4	insertion	0.14	
		800	45.6	insertion		
M-359	C	300	37.1	null-shift		
		500	35.7	insertion	0.25	
		800	36.2	insertion		
M-359	L	300	38.6	null-shift		
		500	36.1	insertion	0.25	
		800	36.9	insertion		
M-359A	S	300	43.3	null-shift		
		500	42.8	insertion	0.12	
		800	44.5	insertion		
M-359	U	300	34.5	null-shift		
		500	34.3	insertion	0.31	
		800	33.4	insertion		
PL-259	A	500			0.036	
	L	500			0.039	
PL-274	L	150	26.9	null-shift		
		500	21.5	insertion		
		800	25	insertion		
		—	22	estimated		
SO-239	A	1				0.007
		10				0.005
		100				0.005
		500			0.15	
		3000	25	insertion		
		—	20	estimated		
SO-239	L	500		insertion	0.13	
		3000	22	insertion		
		—	18	estimated		
SO-239	S	1				0.007
		10				0.004
		100				0.002
		500			0.12	
		3000	25.4	insertion		
		—	23	estimated		
SO-239	W	1				0.05
		10				0.06
		100				0.08
		500	30	insertion	0.31	

Of all types of radio-frequency transmission tests proposed for the socket type SO-239 (Navy type -49194), the method described, used at 3000 megacycles, would probably be capable of yielding the most valuable results. However, in common with any other method, if the test is to evaluate the design or manufacture of the connector, the following precautions must be observed:

Since the results are subject to considerable variation due to the peculiarities of the individual soldering job, and to some variation due to lack of precision in equipment and the personal factor in taking data, the average result of a series of four or five measurements, each after a separate soldering job, should be reported. Great care should be exercised in the design and construction of the line, end socket, etc.

The results of any radio-frequency transmission test of plugs type PL-259 (or Navy type -49195) should be interpreted as being more indicative of the merits of the individual plug-to-cable soldering job than of the design or manufacture of the plug.

Right-angle adaptors, type M-359 (Navy type -49192) and bulkhead plugs, type PL-274, could be tested at a frequency in the neighborhood of from 1000 to 1200 megacycles for greatest precision by this method. However, if many samples are to be tested, it will probably be more convenient to test the right-angle

[1] J. R. Ragazzini, "Transmission lines at UHF," *Elec. Eng.*, vol. 62, p. 165, equation (65); April, 1943.

adaptors at a frequency not over 300 megacycles and the bulkhead plugs at a frequency not over 150 megacycles by the null-shift method, which will be described.

II. The Null-Shift Method of Measurement of Characteristic Impedance of Fittings for Coaxial Connectors

This method is applicable only to adaptors, bulkhead plugs, splicing connectors, etc., and not to plugs and sockets which are attached directly to cables. It depends upon the relationship expressed by (78a) of Section IV

$$Z_c = Z_0\sqrt{\Delta S_s/\Delta S_0}. \qquad (78a)$$

The quantities ΔS_s and ΔS_0 are obtained from four readings on a slotted line for a given sample. ΔS_s is the shift in the null position obtained with the line short-circuited (Fig. 3A) and with the adaptor sample attached and similarly short-circuited (Fig. 3B). ΔS_0 is the shift in the null position obtained with the line open-circuited (Fig. 3C), and with the adaptor attached and open circuited (Fig. 3D). Z_0 is the characteristic impedance of the slotted line and is constant for a given line. The arrangement of the apparatus is shown in Fig. 3. Considerations of the setup for the insertion method apply to this test, except that no cable or indicator calibration is required.

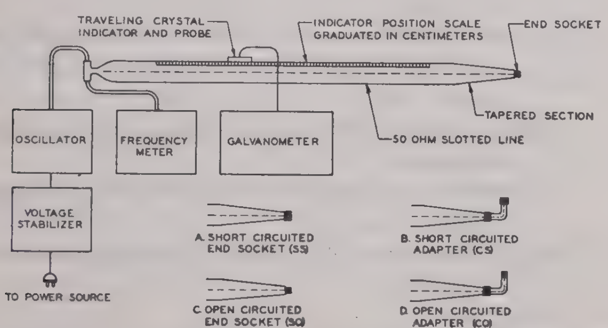


Fig. 3—Arrangement of apparatus for measurement of characteristic impedance of fittings for coaxial connectors by the slotted-line null-shift method.

The coupling between the indicator probe and the inner conductor of the line should be sufficient to permit determination of the null position to within half a millimeter. This degree of coupling may be too great for satisfactory determination of standing-wave ratios, and it may be necessary to provide for variable coupling, to accommodate requirements of the two uses to which the indicator may be put. This is easily done by providing a removable probe extension, which will be attached to the indicator when this null-shift test method is used and which will be removed for measuring standing-wave ratios.

Another difference between the null-shift and insertion methods is that the former employs a considerably lower test frequency. It will be seen from the measurements using the null-shift method on right-angle adaptors, that the results obtained at 500 megacycles are generally higher than those obtained at 300 megacycles. This is particularly noticeable in the longer adaptors, such as the type PL-274, and can be attributed to $\beta_c l_c$ becoming too large at 500 megacycles for strict justification of the approximation made in Section IV, (69). For this reason, measurements on right-angle adaptors (type M-359) should be made at not over 300 megacycles and measurements on bulkhead plugs (type PL-274) should be made at 150 megacycles or below. This should be satisfactory for all purposes since theoretically Z_c is not a function of frequency. Further argument for the use of a lower test frequency in production testing is the greater ease of obtaining satisfactory radio-frequency oscillators at lower frequencies.

A. *Evaluation of the Null-Shift Method*

The null-shift method of evaluating adaptors appears to have the following advantages over the insertion method: (1) No indicator calibration is necessary, so that time is saved and a source of error eliminated. (2) There is no rigid requirement on the voltage stability of the radio-frequency source, which permits the use of some oscillators found unsuitable where constant output is required. (3) Great care in minimizing the loading effect of the indicator is unnecessary, and relatively tight coupling to the line can be made since readings are taken at nulls, where negligible power is taken. (4) After the equipment has been set up, successive measurements can be made very rapidly. (5) Measurements can be made on adaptors designed for impedances other than the characteristic impedance of the test equipment.

III. Power Factor of Connectors

One method of measuring the power factor of the dielectric of connectors employs the Q meter. Readings of C_1 and Q_1 are taken with only a shielded inductor in place, and readings of C_2 and Q_2 are obtained after connecting the test specimen across the "COND." terminals of the Q meter. The formula for power factor is then[2]

$$P = C_1(Q_1 - Q_2)/(C_1 - C_2)Q_1Q_2.$$

The total capacitance is given[2] by the expression $C_p = C_1 - C_2$.

It has been found in practice that best results are obtained with an inductor having the highest possible Q_1, which, with the specimen connected, will resonate a value of C_2 near the lower limit of the capacitance dial range.

IV. Mathematical Derivation of Theory

A. *Definitions of Symbols Used*

A $= \sqrt{(R^2+X^2-Z_0^2)^2+4X^2Z_0^2}$; introduced for convenience

α = attenuation constant of slotted line in nepers per centimeter

[2] "Instructions and Manual of Radio Frequency Measurements for Q Meters Type 100-A, Type 160-A, and Type 170-A," Boonton Radio Corporation, Boonton, N. J., equation (40), page 16, and equation (26b), page 7, June, 1941.

α_c = attenuation constant of connector in nepers per centimeter

β = wavelength constant of slotted line in radians per centimeter $= 2\pi/\lambda = 2\pi f/c$

β_c = wavelength constant of connector in radians per centimeter

β_s = wavelength constant of slotted-line end socket in radians per centimeter

C = capacitance of connector in farads per centimeter

C_A = capacitance of connector in farads per centimeter if dielectric were replaced by air

c = velocity of light in centimeters per second

γ $= \alpha + j\beta$

γ_c $= \alpha_c + j\beta_c$

D $= Z_c/Z_0$; introduced for convenience

δ $= \tan^{-1}(R^2 + X^2 - Z_0^2)/2XZ_0$, introduced for convenience

E_r = voltage at receiving end of slotted line

E_s $= E_s\underline{/\phi}$ voltage on line at distance l from connector

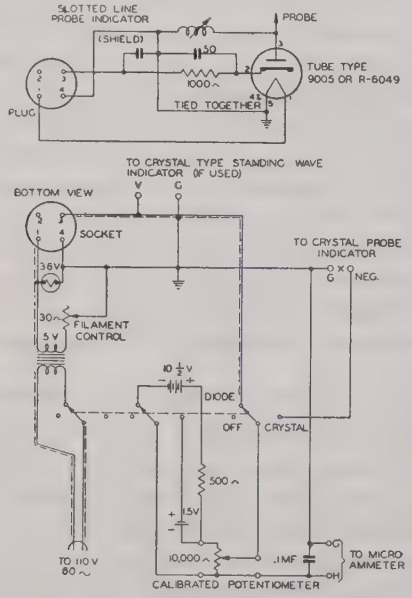


Fig. 4—Suggested circuit for precision slideback voltmeter for general slotted-line use.

$E_{\min}$ = minimum value of E_s along the line; at distance l_m from connector (See Fig. 4.)

$E_{\max}$ = maximum value of E_s along the line; at a distance $l_{\max}$ from connector (See Fig. 4.)

ϵ = dielectric constant of connector $= C/C_A$

ζ = an upper limit of the relative error in Z_c

η = maximum relative indicator error

f = test frequency in cycles per second

f_1 = an arbitrarily chosen value of f

f_0 = value of f resulting in least error in determining Z_c of a given connector

H $= \tan(\pi f/2f_0)$ introduced for convenience

I_r = current at input of connector terminated in Z_0

j $= \sqrt{-1}$

L = inductance of a connector in henries per centimeter

l = distance in centimeters along the line from connector to point of voltage E_s

l_c = electrical length of connector in centimeters

l_{co} $= l_m$ for line terminated in open-circuited connector

l_{cs} $= l_m$ for line terminated in a short-circuited connector

l_m = distance in centimeters along the line from connector to point of minimum voltage

$l_{\max}$ = distance in centimeters along the line from connector to point of maximum voltage

l_s = electrical length of end socket in centimeters

l_{so} $= l_m$ for open-circuited end socket

l_{ss} $= l_m$ for condition of short-circuited line $(R = X = 0)$

l_{st} $= l_m$ for condition of connector terminated in Z_0

λ = wavelength in centimeters $= 2\pi/\beta$

r $= E_{\min}/E_{\max}$; (real and between 0 and 1)

R = resistance of slotted-line load in ohms (general case)

R_{in} = input resistance in ohms of connector when terminated in Z_0, in ohms

ΔS = shift of point of minimum voltage on slotted line when termination is changed from short circuit to connector terminated in Z_0; $= l_{st} - l_{ss}$

ΔS_{os} = line null shift in centimeters due to short-circuiting the end socket $= l_{so} - l_{ss}$

ΔS_o = line null shift in centimeters due to changing line load from open circuit to open-circuited fitting $= l_{co} - l_{so}$

ΔS_s = line null shift in centimeters due to changing line load from short circuit to short-circuited fitting $= l_{cs} - l_{ss}$

ω $= 2\pi f$ = angular velocity of test frequency in radians per second

X = reactance termination of slotted line, in ohms (general case)

X_{in} = input reactance of connector when terminated in Z_o, in ohms

Z $= R + jX$

Z_c = characteristic impedance of fitting, in ohms

Z_{in} $= R_{in} + jX_{in}$

Z_o = characteristic impedance of slotted line, in ohms

Z_s = characteristic impedance of end socket, in ohms

Z_{so} = input impedance of open-circuited end socket in ohms $\cong jX_{so}$

Z_{ss} = input impedance of short-circuited end socket in ohms $\cong jX_{ss}$

Z_{co} = input impedance of open-circuited fitting, in ohms $\cong jX_{co}$

Z_{cs} = input impedance of short-circuited fitting, in ohms $\cong jX_{cs}$

B. *The Impedance Termination in Terms of r and $\beta 1_m$*

A general equation has been formulated[3] which may be applied to the slotted line:

$$E_s = E_r \cosh \gamma l + I_r Z_0 \sinh \gamma l. \tag{1}$$

[3] W. L. Everitt, "Communication Engineering," McGraw-Hill Book Co., New York, N. Y., second edition, 1937, p. 159, equation (48).

Since α may be considered negligible, we have:

$$E_s \cos\phi + jE_s \sin\phi = I_r[R\cos\beta l + j(X\cos\beta l + Z_0 \sin\beta l)]. \quad (2)$$

The norm of (2) is

$$E_s^2 = I_r^2[R^2\cos^2\beta l + (X\cos\beta l + Z_0\sin\beta l)^2]. \quad (3)$$

from which

$$E_s^2 = I_r^2[(R^2+X^2+Z_0^2) + (R^2+X^2-Z_0^2)\cos 2\beta l + 2XZ_0 \sin 2\beta l]/2 \quad (3a)$$

which may be presented in the form

$$E_s^2 = I_r^2[(R^2+X^2+Z_0^2) + A\sin(2\beta l+\delta)]/2 \quad (3b)$$

where $A = \sqrt{(R^2+X^2-Z_0^2)^2+4X^2Z_0^2}$
and $\delta = \tan^{-1}[(R^2+X^2-Z_0^2)/2XZ_0]$.

We wish to obtain $R+jX$ in terms of $r=E_{\min}/E_{\max}$ and βl_m. To find $E_{\min}$ and $E_{\max}$, we must have

$$dE_s^2/d\beta l = I_r^2 A\cos(2\beta l+\delta) = 0, \quad (4)$$

so that

$$2\beta l + \delta = (2n+1)\pi/2, \quad n = 0, 1, 2, 3, \cdots, \quad (5)$$

and, substituting in (3b),

$$\left.\begin{aligned} E_{\min}^2 &= (I_r^2/2)[(R^2+X^2+Z_0^2) - A] \\ E_{\max}^2 &= (I_r^2/2)[(R^2+X^2+Z_0^2) + A] \end{aligned}\right\} \quad (6)$$

are given by odd and even values of n respectively, so that, taking $n=2m+1$ and $n=2m$ in (5) we have

$$\left.\begin{aligned} 2\beta l_m+\delta &= (4m+3)\pi/2 \\ 2\beta l_{\max}+\delta &= (4m+1)\pi/2 \end{aligned}\right\} \text{ where } m=0, 1, 2, 3, \cdots. \quad (7)$$

Thus, we see that $\beta l_{\max}$ differs from βl_m by an odd multiple of $\pi/2$.

Since $\delta=(4m+3)\pi/2-2\beta l_m$, we have $\sin\delta=-\cos 2\beta l_m$ and $\cos\delta=-\sin 2\beta l_m$, and consequently

$$\tan\beta l_m = \begin{cases} \dfrac{1-\cos 2\beta l_m}{\sin 2\beta l_m} = \dfrac{1+\sin\delta}{-\cos\delta} = -\dfrac{A+R^2+X^2-Z_0^2}{2XZ_0} & (8a) \\[2ex] \dfrac{\sin 2\beta l_m}{1+\cos 2\beta l_m} = \dfrac{-\cos\delta}{1-\sin\delta} = \dfrac{-2XZ_0}{A-R^2-X^2+Z_0^2}. & (8b) \end{cases}$$

Now, from (6)

$$r^2 = \frac{R^2+X^2+Z_0^2-A}{R^2+X^2+Z_0^2+A} = \begin{cases} [R^2+X^2+Z_0^2-A]^2/4R^2Z_0^2 & (9a) \\ 4R^2Z_0^2/[R^2+X^2+Z_0^2+A]^2. & (9b) \end{cases}$$

Thus

$$r = \begin{cases} (R^2+X^2+Z_0^2-A)/2RZ_0 & (10a) \\ 2RZ_0/(R^2+X^2+Z_0^2+A). & (10b) \end{cases}$$

The simplest way to find R and X is to obtain two linear equations by eliminating $A+R^2+X^2$ between (8a) and (10b), and $-A+R^2+X^2$ between (8b) and (10a); upon dividing the results through by $2Z_0$, multiplying the first through by $r\cos\beta l_m$, and the second by $\sin\beta l_m$, we obtain

$$Z_0 r\cos\beta l_m = rX\sin\beta l_m + R\cos\beta l_m \quad (11a)$$
$$Z_0\sin\beta l_m = -X\cos\beta l_m + Rr\sin\beta l_m. \quad (11b)$$

Multiplying (11b) by $-j$ and adding to (11a) yields, upon factoring

$$R+jX = Z_0\frac{r\cos\beta l_m - j\sin\beta l_m}{\cos\beta l_m - jr\sin\beta l_m} = \frac{Z_0(r-j\tan\beta l_m)}{1-jr\tan\beta l_m}. \quad (12)$$

We can obtain formulas for R and X either by solving (11a) and (11b) simultaneously, or by equating real and imaginary parts of (12).

$$R = Z_0 r/(\cos^2\beta l_m + r^2\sin^2\beta l_m) \quad (13a)$$
$$X = -Z_0(1-r^2)\sin\beta l_m\cos\beta l_m/(\cos^2\beta l_m + r^2\sin\beta l_m). \quad (13b)$$

From these it follows that

$$A = Z_0^2(1-r^2)/(\cos^2\beta l_m + r^2\sin^2\beta l_m). \quad (13c)$$

C. *Application to the Slotted Line*

To find βl_{ss}, we must apply the conditions $R=X=0$. Under these conditions, $A=Z_0^2$ and, from the definition of δ, when $R=X=0$,

$$\sin\delta = (R^2+X^2-Z_0^2)/A = -1. \quad (14)$$

Hence δ differs from $3\pi/2$ by a multiple of 2π, so that, from (7), we obtain

$$\beta l_{ss} = m\pi, \quad \text{where} \quad m = 0, 1, 2, 3, \cdots. \quad (15)$$

It therefore follows that

$$\beta\Delta S = \beta l_{st} - \beta l_{ss} = \beta l_{st} - m\pi, \quad (16)$$

and consequently

$$\tan\beta\Delta S = \tan\beta l_{st}. \quad (17)$$

We are now ready to apply (12) to the slotted line terminated in a connector which is in turn terminated in a load of characteristic impedance, Z_0. Substituting βl_{st} for βl_m in (12), and using (17), we have

$$R_{in} + jX_{in} = Z_0(r - j\tan\beta\Delta S)/(1 - jr\tan\beta\Delta S). \quad (18)$$

D. *The Characteristic Impedance of the Connector*

Using[4] Everitt's (44) for the input impedance of the connector, we have

$$\begin{aligned} R_{in} + jX_{in} &= Z_c\frac{Z_0\cosh j\beta_c l_c + Z_c\sinh j\beta_c l_c}{Z_c\cosh j\beta_c l_c + Z_0\sinh j\beta_c l_c} \\ &= Z_c\frac{Z_0\cos\beta_c l_c + jZ_c\sin\beta_c l_c}{Z_c\cos\beta_c l_c + jZ_0\sin\beta_c l_c} \\ &= Z_c\frac{Z_0 + jZ_c\tan\beta_c l_c}{Z_c + jZ_0\tan\beta_c l_c}. \end{aligned} \quad (19)$$

Equating (18) and (19), and letting $Z_c/Z_0 = D$, we have

$$D\frac{1+jD\tan\beta_c l_c}{D+j\tan\beta_c l_c} = \frac{r-j\tan\beta\Delta S}{1-jr\tan\beta\Delta S}. \quad (20)$$

Upon cross-multiplying, and equating the real and imaginary components respectively, we obtain

$$\left.\begin{aligned} D(1+rD\tan\beta\Delta S\tan\beta_c l_c) &= rD + \tan\beta\Delta S\tan\beta_c l_c \\ D(D\tan\beta_c l_c - r\tan\beta\Delta S) &= r\tan\beta_c l_c - D\tan\beta\Delta S \end{aligned}\right\}. \quad (21)$$

It follows that

$$\left.\begin{aligned} \tan\beta\Delta S\tan\beta_c l_c &= D(1-r)/(1-rD^2) \\ D(1-r)\tan\beta\Delta S &= (r-D^2)\tan\beta_c l_c \end{aligned}\right\} \quad (22)$$

Thus

$$\tan^2\beta\Delta S = (r-D^2)/(1-rD^2) \quad (23a)$$
$$\tan^2\beta_c l_c = D^2(1-r)^2/(1-rD^2)(r-D^2). \quad (23b)$$

Equation (23a) yields $(1-r\tan^2\beta\Delta S)D^2 = r - \tan^2\beta\Delta S$, so that

$$Z_c = Z_0\sqrt{(r-\tan^2\beta\Delta S)/(1-r\tan^2\beta\Delta S)} \quad (24)$$

[4] See p. 158 of footnote reference 3.

$$Z_c = Z_0\sqrt{[r-\tan^2(2\pi\Delta S/\lambda)]/[1-r\tan^2(2\pi\Delta S/\lambda)]}. \quad (24a)$$

For $Z_0=50$ ohms, these curves have been plotted in Fig. 2.

From (23b) and (68), which will be derived later in this section, we have

$$\tan^2\beta l_{cs} = D^2\tan^2\beta_c l_c = D^4(1-r)^2/(1-rD^2)(r-D^2). \quad (25)$$

Just as in (15), (16) and (17), we have

$$\beta\Delta S_s = \beta l_{cs} - m\pi \quad (26)$$

$$\tan\beta\Delta S_s = \tan\beta l_{cs}, \text{ so that, finally,} \quad (27)$$

$$\tan^2\beta\Delta S_s = D^4(1-r)^2/(1-rD^2)(r-D^2). \quad (28)$$

Hence, by definition,

$$\tan^2(2\pi\Delta S_s/\lambda) = (Z_c/Z_0)^4(1-r)^2/[(1-r(Z_c^2/Z_0^2)][(r-(Z_c^2/Z_0^2)]. \quad (28a)$$

E. *Ranges of Variation; Errors Due to Error in Indicator Reading; Regions of Optimum Applicability*

From (23a) we have

$$(1-rD^2)\tan^2\beta\Delta S - (r-D^2) = 0. \quad (29)$$

For convenience, let us call the left-hand side of this equation $f(r, D, \beta\Delta S)$; then, using (23a) after differentiating,

$$\partial f/\partial r = -(1+D^2\tan^2\beta\Delta S) = -(1-D^4)/(1-rD^2)$$
$$\partial f/\partial D = 2D(1-r\tan^2\beta\Delta S) = 2D(1-r^2)/(1-rD^2)$$
$$\partial f/\partial\beta\Delta S = 2(1-rD^2)\tan\beta\Delta S\sec^2\beta\Delta S = \pm 2(1+r)(1-D^2)\sqrt{(r-D^2)/(1-rD^2)}$$

and in the last equation the sign is that of $\tan\beta\Delta S$; i.e.,

$$\left\{\begin{array}{l}\text{positive for } n\pi \leqq \beta\Delta S < (2n+1)\pi/2 \\ \text{negative for } (2n+1)\pi/2 < \beta\Delta S \leqq (n+1)\pi\end{array}\right\} \quad n = 0, \pm 1, \pm 2, \cdots.$$

We therefore have, as total differential of (29), after multiplying through by $1-rD^2$ and noting that

$$(1-rD^2)\sqrt{(r-D)/(1-rD^2)} = -\sqrt{(r-D^2)(1-rD^2)}$$

when $(1-rD^2)<0$,

$$-(1-D^4)dr + 2D(1-r^2)dD \pm 2(1+r)(1-D^2)\sqrt{(r-D^2)(1-rD^2)}\,d\beta\Delta S = 0 \quad (30)$$

where the sign is now

$$\left\{\begin{array}{l}\text{positive when } 1-rD^2 \text{ has the same sign as } \tan\beta\Delta S \\ \text{negative when } 1-rD^2 \text{ has sign opposite to that of } \tan\beta\Delta S\end{array}\right\}. \quad (31)$$

Thus we have

$$dD = [(1-D^4)/2D(1-r^2)]dr - (\pm 1)(1-D^2)[\sqrt{(r-D^2)(1-rD^2)}/D(1-r)]d\beta\Delta S \quad (32)$$

where the sign is determined as in (31).

We are now in a position to describe the types and ranges of variation in (23a).

(1) $r-D^2$ and $1-rD^2$ must have the same sign. Thus, either $D^2<r$ and $D^2<(1/r)$, or $D^2>r$ and $D^2>(1/r)$; for the region in which we are interested, namely $0\leqq r\leqq 1$, $D\geqq 0$, we therefore have either $D^2<r$ or $D^2>(1/r)$. This explains why all the curves in Fig. 2 lie either under the parabola $D^2=r$ or over the curve of hyperbolic type $D^2=(1/r)$.

(2) $D=0$ when $\tan^2\beta\Delta S=r$, and at these points the curves are perpendicular to the r axis, for (30) yields, when $d\beta\Delta S=0$,

$$dD/dr = (1-D^4)/2D(1-r^2) \quad (33)$$

which approaches infinity as D approaches 0. Also, (33) shows that, for a fixed $\beta\Delta S$, D increases with r when $D<1$ and decreases as r increases for $D>1$, assuming, of course, that $r<1$, $(\beta=2\pi/\lambda)$.

(3) For $\beta\Delta S=n\pi$; i.e., for $\Delta S=(n/2)\lambda$, $r=D^2$; for $\beta\Delta S=n\pi+\pi/4$; i.e., for $\Delta S=(n/2+1/8)\lambda$, $r=1$; for $\beta\Delta S=n\pi+\pi/2$; i.e., for $\Delta S=(n/2+1/4)\lambda$, $r=1/D^2$.

(4) For constant D, taking $dD=0$ in (30), and using (23b) we have

$$\frac{dr}{d\beta\Delta S} = \frac{\pm 2(1+r)\sqrt{(r-D^2)(1-rD^2)}}{1+D^2} = \frac{2D(1-r^2)}{(1+D^2)\tan\beta_c l_c} \quad (34)$$

where the sign is determined by (31). This means that where $r<1$,

$$r \text{ increases with } \beta\Delta S \text{ for } \left\{\begin{array}{l}D>1,\ \tan\beta\Delta S<0 \\ D<1,\ \tan\beta\Delta S>0\end{array}\right\}$$

$$r \text{ decreases with } \beta\Delta S \text{ for } \left\{\begin{array}{l}D>1,\ \tan\beta\Delta S>0 \\ D<1,\ \tan\beta\Delta S<0\end{array}\right\}.$$

For example, in Fig. 2

$$r \text{ increases with } \Delta S \text{ for } \left\{\begin{array}{l}Z_c>50,\ -\lambda/4\leqq\Delta S\leqq-\lambda/8 \\ Z_c<50,\ 0\leqq\Delta S\leqq\lambda/8\end{array}\right\}$$

$$r \text{ decreases with } \Delta S \text{ for } \left\{\begin{array}{l}Z_c>50,\ \lambda/8\leqq\Delta S\leqq\lambda/4 \\ Z_c<50,\ -\lambda/8\leqq\Delta S\leqq 0\end{array}\right\}. \quad (35)$$

Having discussed the variation in (23a), we come now to a discussion of the effect of errors on the final result. Equation (32) shows the effect of an error in r and in $\beta\Delta S$; both of these are due to errors in the reading. We will find limits for dr and $d\beta\Delta S$ in terms of the error in indicator reading η and substitute back in (32).

(5) Since we have $r=E_{\min}/E_{\max}$ it follows that $\log r=\log E_{\min}-\log E_{\max}$, so that

$$dr/r = dE_{\min}/E_{\min} - dE_{\max}/E_{\max}. \text{ Hence} \quad (36)$$

$$|dr/r| < 2\eta, \quad \text{and} \quad |dr| < 2\eta r. \quad (36a)$$

(6) Since $E_{\min}$ and $E_{\max}$ have been found by varying $\beta\Delta S$, (4) shows that we must go to second order terms in finding their errors in terms of $d\beta\Delta S$. From (4) we have, noting (16)

$$dE_s/d\beta\Delta S = I_r^2A\cos(2\beta\Delta S+\delta)/2E_s, \text{ so that}$$

$$d^2E_s/d(\beta\Delta S)^2 = [-4E_sI_r^2A\sin(2\beta\Delta S+\delta) - 2I_r^2A\cos(2\beta\Delta S+\delta)dE_s/d\beta\Delta S]/4E_s^2$$

and substituting from (7) yields

$$[d^2E_s/(d\beta\Delta S)^2]_{\min} = I_r^2A/E_{\min} \quad (37a)$$

$$[d^2E_s/(d\beta\Delta S)^2]_{\max} = -I_r^2A/E_{\max}. \quad (37b)$$

Going to the third term of the Taylor expansion, we have

$$E_s(\beta\Delta S+d\beta\Delta S) = E_s(\beta\Delta S) + (dE_s/d\beta\Delta S)d\beta\Delta S + 1/2[d^2E_s/(d\beta\Delta S)^2](d\beta\Delta S)^2.$$

Consequently, using (4), (37a), and (37b), we have

$$\Delta E_{\min} \doteq (I_r^2 A/2E_{\min})(d\beta\Delta S)^2 \quad (38a)$$

$$\Delta E_{\max} \doteq (-I_r^2 A/2E_{\max})(d\beta\Delta S)^2. \quad (38b)$$

It follows from (6) that

$$\Delta E_{\min}/E_{\min} \doteq [A/(R^2+X^2+Z_0^2-A)](d\beta\Delta S)^2 \quad (39a)$$

$$\Delta E_{\max}/E_{\max} \doteq [-A/(R^2+X^2+Z_0^2+A)](d\beta\Delta S)^2. \quad (39b)$$

Since, from (10a) and (10b),

$R^2+X^2+Z_0^2-A=2rZ_0$; and $R^2+X^2+Z_0^2+A=(2/r)RZ_0$,

we may substitute from (13a) and (13c) to obtain

$$\Delta E_{\min}/E_{\min} \doteq [(1-r^2)/2r^2](d\beta\Delta S)^2 \quad (40a)$$

$$\Delta E_{\max}/E_{\max} \doteq -[(1-r^2)/2](d\beta\Delta S)^2 \quad (40b)$$

so that

$$\Delta E_{\min}/E_{\min}-\Delta E_{\max}/E_{\max} = [(1-r^4)/2r^2](d\beta\Delta S)^2. \quad (40c)$$

As in Section IV, E(5), we therefore have

$$[(1-r^4)/2r^2](d\beta\Delta S)^2 < 2\eta, \quad (41)$$

and

$$|d\beta\Delta S| < (2r/\sqrt{1-r^4})\sqrt{\eta}. \quad (41a)$$

(7) We now find an estimated maximum relative error in D, which we will call ζ, by transforming (32) into the inequality

$$\left|\frac{dD}{D}\right| \leq \left|\frac{1-D^4}{2D^2(1-r^2)}\right| |dr| + \left|\frac{(1-D^2)\sqrt{(r-D^2)(1-rD^2)}}{D^2(1-r)}\right| |d\beta\Delta S|$$

and using inequalities (36a) and (41a)

$$\zeta = \left|\frac{1}{D^2}-D^2\right|\left|\frac{r}{1-r^2}\right|\eta + 2\left|\frac{1}{D^2}-1\right|\left|\frac{r}{(1-r)\sqrt{1-r^4}}\right|\sqrt{(r-D^2)(1-rD^2)}\,\sqrt{\eta}. \quad (42)$$

Since $D=Z_c/Z_0$, and since the relative error in Z_c is the same as that of D

$$\zeta = \left|\frac{Z_0^2}{Z_c^2}-\frac{Z_c^2}{Z_0^2}\right|\left|\frac{r}{1-r^2}\right|\eta + 2\left|\frac{Z_0^2}{Z_c^2}-1\right|\left|\frac{r}{(1-r)\sqrt{1-r^4}}\right| \sqrt{\left(r-\frac{Z_c^2}{Z_0^2}\right)\left(1-r\frac{Z_c^2}{Z_0^2}\right)\eta}. \quad (42a)$$

(8) The procedure in calculating this upper limit of error may be co-ordinated with the measurements as follows:

First, readings are taken for $E_{\min}$, $E_{\max}$, and ΔS. The $r=E_{\min}/E_{\max}$ is calculated, and r and ΔS are used to find Z_c either from (24) or Fig. 2. These values of r and Z_c are used, instead of their true values, in (42a). The first term gives an upper limit of error due to wrong r; the second yields an upper limit of error due to wrong ΔS.

It is obvious from (42a) that the errors are very large when Z_c is close to zero, and also when r is close to one. These are the regions in which the curves in Fig. 2 become steep. The errors become smaller with decreasing r; the smallest r for a given Z_c is on the limiting curves of Fig. 2.

$$\Delta S = 0; \qquad Z_c^2 = rZ_0^2$$

and

$$\Delta S = \pm\lambda/4; \qquad Z_0^2 = rZ_c^2.$$

Hence, the radical in the second term of (42a) approaches zero as we approach these curves for a fixed Z_c. Close to these curves, therefore, the second error term becomes negligible and the first error term gets increasingly smaller; in fact, substituting

$$r = Z_c^2/Z_0^2 \quad \text{or} \quad r = Z_0^2/Z_c^2$$

we see that on these curves $\zeta=\eta$. It therefore follows that for constant Z_c, as ΔS approaches 0 or $\pm\lambda/4$, the error decreases to that of the indicator.

F. *Variation with Frequency; Choice of Optimum Frequency*

(1) From (23b) it is obvious that $D^2=r$ or $D^2=1/r$ if and only if $\beta_c l_c=\pi/2+n\pi$. Combining this with remarks in the preceding paragraph, we have

$$\begin{cases} \text{as } \beta_c l_c \text{ approaches } \pi/2+n\pi, \\ r \text{ decreases to } Z_c^2/Z_0^2 \text{ or to } Z_0^2/Z_c^2, \text{ and} \\ \Delta S \text{ approaches either 0 or } \pm\lambda/4. \end{cases} \quad (43)$$

Furthermore, if $0 \leqq \beta_c l_c < \pi/2$, tan $\beta_c l_c$ is positive, and the second equation of (22) shows that tan $\beta\Delta S$ has the same sign as $r-D^2$. Thus, in the range $0 \leqq r \leqq 1$ in which we are interested we have, since (34) shows that r must increase, from (35)

if $0 \leqq \beta_c l_c < \pi/2$, then

$$\begin{cases} -\lambda/4 \leqq \Delta S \leqq -\lambda/8 \text{ for } Z_c > Z_o \\ 0 \pm \Delta S \pm \lambda/8 \text{ for } Z_c < Z_o. \end{cases} \quad (44)$$

(2) In view of the preceding paragraphs, we see that the error in obtaining Z_c will be minimized by so choosing the operating frequency that ΔS is close to 0 or $\pm\pi/4$, in which case $\beta_c l_c$ is close to $(n+1/2)\pi$. Both β and β_c vary with the operating frequency in a manner that we will discuss in a moment. But, whereas ΔS also varies with the operating frequency, l_c does not. We will therefore use (23b) rather than (23a) to help us find the best operating frequency, f_0, for which $\beta_c l_c=\pi/2$.

From Everitt,[6] we have

$$\omega/\beta = 1/\sqrt{LC_A} \text{ and } \omega/\beta_c = 1/\sqrt{LC}, \text{ so that}$$

$$\beta c/\beta = \sqrt{C/C_A} = \sqrt{\epsilon}.$$

But for f_0 we have $\beta_c l_c=\pi/2$ and $\beta=2\pi f_o/c$, hence $\sqrt{\epsilon}=\beta_c/\beta=c/4l_c f_0$, so that

$$f_0 = c/4l_c\sqrt{\epsilon} = 7.5\cdot 10^9/l_c\sqrt{\epsilon}. \quad (45)$$

Since, in general, $\beta_c=2\pi f\sqrt{\epsilon}/c$ so that β_c is proportional to f, and since l_c is independent of f, we have $\beta_c l_c$ proportional to f, so that

$$f_1/f_0 = (\beta_c l_c)_1/(\beta_c l_c)_0 = 2\beta_c l_c/\pi \quad (46)$$

and

$$\beta_c l_c = (\pi/2)(f_1/f_0). \quad (46a)$$

Substituting (46a) in (23b) we obtain

$$\tan^2\left(\frac{\pi}{2}\frac{f_1}{f_0}\right) = \frac{D^2(1-r)^2}{(1-rD^2)(r-D^2)} = \frac{Z_c^2Z_0^2(1-r)^2}{(Z_0^2-rZ_c^2)(rZ_0^2-Z_c^2)}. \quad (47)$$

[6] See p. 118 of footnote reference 3.

In terms of ΔS rather than D, this becomes

$$\tan^2\left(\frac{\pi}{2}\frac{f_1}{f_0}\right)=\frac{(r-\tan^2\beta\Delta S)(1-r\tan^2\beta\Delta S)}{(1-r)^2\tan^2\beta\Delta S}. \quad (47a)$$

These curves are traced as dotted lines in Fig. 2 for various values of f_1/f_0. For calculation purposes it is perhaps best to solve for r; let $H=\tan[(\pi/2)(f_1/f_0)]$.

$$r=1+\frac{H(1-D^2)}{2D^2(1+H^2)}\left[H(1-D^2)\pm\sqrt{4D^2+H^2(1+D^2)^2}\right]$$
$$=1+\frac{Z_0^2-Z_c^2}{2Z_0^2Z_c^2}\cdot\frac{H}{1+H^2}\left[H(Z_0^2-Z_c^2)\pm\sqrt{4Z_0^2Z_c^2+H^2(Z_0^2-Z_c^2)^2}\right]. \quad (48)$$

Since $0<r<1$, the sign will be $+$ or $-$ as $Z_c>Z_o$ or $Z_c<Z_o$, respectively.

To find a good operating frequency, (45) is used as an approximation, f_1 is chosen definitely less than this approximation, measurements are made, and $(r, \Delta S)$ spotted. The dotted line in Fig. (2) on which this point would lie will give a value of f_0/f_1, and a better approximation of f_0 is found. It is the application of (45) to approximate values of l_c and ϵ that makes (45) a poorer approximation to f_0 than the one obtained by the process just described.

G. *Theory of Tailoring of the Slotted-Line End Socket*

Converting[4] Everitt's (44a)

$$Z_{so}=Z_s\frac{(1/j\omega C)\cos\beta_s l_s+jZ_s\sin\beta_s l_s}{Z_s\cos\beta_s l_s+(\sin\beta_s l_s)/\omega c} \quad (49)$$
$$X_{so}=-Z_s\cot(\beta_s l_s+\tan^{-1}Z_{s\omega}C) \quad (49a)$$

and from (45) of the same reference, we obtain

$$Z_{ss}=jZ_s\tan\beta_s l_s \quad (50)$$
$$X_{ss}=Z_s\tan\beta_s l_s. \quad (50a)$$

From (8a), we obtain, in the case of a pure reactance termination, $(R=0)$,

$$X=-Z_0\tan\beta l_m. \quad (51)$$

Using the open-circuit condition, where the value of l_m becomes l_{so} and the value of X becomes X_{so}, equations (1a) and (3) give

$$\beta l_{so}=\tan^{-1}[(Z_s/Z_0)\cot(\beta_s l_s+\tan^{-1}Z_s\omega C)]$$
$$=\tan^{-1}[(Z_s/Z_0)\tan(\pi/2-\beta_s l_s-\tan^{-1}Z_s\omega C)]. \quad (52)$$

For the short-circuited condition, where the value of X becomes X_{ss}, combining (50a) and (51), we obtain

$$\beta l_{ss}=-\tan^{-1}[(Z_s/Z_0)\tan\beta_s l_s]. \quad (53)$$

Also

$$\Delta S_{os}=l_{so}-l_{ss}=(1/\beta)\tan^{-1}[(Z_s/Z_0)\tan(\pi/2-\beta_s l_s-\tan^{-1}Z_s\omega C)]+(1/\beta)\tan^{-1}[(Z_s/Z_0)\tan\beta_s l_s] \quad (54)$$
$$\cot\beta\Delta S_{os}=[(Z_0/Z_s)\tan\beta_s l_s+Z_0\omega C-(Z_s/Z_0)(1-\tan\beta_s l_s Z_s\omega C)\tan\beta_s l_s]\cos^2\beta_s l_s. \quad (54a)$$

Now we wish to show that ΔS_{os} increases as Z_s increases for conditions encountered in this method of measurement, namely

$$0<\beta_s l_s<\pi/4 \quad (55)$$
$$Z_0/2<Z_s<2Z_0 \quad (56)$$
$$Z_0\cong 50 \quad (57)$$
$$0<\omega C<10^{-3}. \quad (58)$$

To do this, we first show that $(d\cot\beta\Delta S_{os})/dZ_s$ does not change sign at any point within these limits.

$$\frac{d\cot\beta\Delta S_{os}}{dZ_s}=\frac{\sin 2\beta_s l_s}{2Z_s^2Z_0}[Z_s^3(2\tan\beta_s l_s\omega C)-Z_s^2-Z_0^2]. \quad (59)$$

Then we obtain

$$(d\cot\beta\Delta S)/dZ_s=[(\sin 2\beta_s l_s)/(2Z_s^2Z_0)]F(Z_s) \quad (59a)$$

by letting

$$F(Z_s)=Z_s^3(2\tan\beta_s l_s\omega C)-Z_s^2-Z_0^2. \quad (60)$$
$$F'(Z_s)=2Z_s(3\omega CZ_s\tan\beta_s l_s-1). \quad (61)$$

Therefore,

$$F'(Z_s)=0 \text{ when } Z_s=0, \quad (62)$$
in which case $F(Z_s)=-Z_0^2$, or
$$Z_s=1/3\omega C\tan\beta_s l_s, \quad (63)$$
in which case $F(Z_s)=-1/(27\omega C\tan\beta_s l_s)-Z_0^2$.

From (55) and (58) it is seen that the minimum value of Z_s to be expected at point (63) is 333, and Z_s is zero at point (62), so that from (56) and (57) it appears that Z_s must lie between these points. Since (60) is a polynomial in Z_s, and therefore continuous everywhere, a reversal of slope of $F(Z_s)$ can occur only where $F'(Z_s)=0$, namely the points of (62) and (63). Also, since $F(Z_s)$ is negative at these points, and there is no reversal of slope in the region between them, $F(Z_s)$ must be negative over this entire region. Applying (55), (56), and (57) to (59a) shows $d(\cot\beta\Delta S)/dZ_s$ to be of the same sign as $F(Z_s)$ over this region, so that $d(\cot\beta\Delta S)/dZ_s$ is negative and $\cot\beta\Delta S$ decreases as Z_s increases. From (54a) it is seen that $\beta\Delta S$, being an arccotangent, has an infinite number of values, each differing by $n\pi$, where n is a whole integer. For these measurements, only the value of $\beta\Delta S$ between zero and π is used, so that $\beta\Delta S$ increases as $\cot\beta\Delta S$ decreases, and therefore, since β is a positive constant, ΔS increases as Z_s increases. From transmission line theory,

$$\beta=2\pi/\lambda=\omega/c \quad (64)$$

and for the condition of $Z_s=Z_0$, (54) becomes, for practical conditions,

$$\Delta S_{os}=\frac{\pi/2-\tan^{-1}Z_0\omega C}{\beta}\cong\frac{\pi/2-Z_0\omega C}{\beta}=\frac{\lambda}{4}-Z_0C_c. \quad (65)$$

For the case of $Z_0=50$, this becomes,

$$\Delta S_{os}=\lambda/4-1.5C_{\mu\mu f}. \quad (66)$$

H. *Theory of Measurement of Characteristic Impedance of of Fitting*

Converting[4] Everitt's (45) to the terminology of this report, we obtain

$$Z_{cs}=jZ_c\tan\beta_c l_c \quad (67)$$
$$X_{cs}=Z_c\tan\beta_c l_c. \quad (67a)$$

For the case of a short circuited fitting, X becomes X_{cs}, and l_m becomes l_{cs}, so that combining (51) and (57a) yields

$$-Z_0\tan\beta l_{cs}=Z_c\tan\beta_c l_c. \quad (68)$$

For conditions encountered in practice,

$$\beta_c l_c\ll\pi/4, \quad (69)$$

and

$$Z_c\cong Z_0. \quad (70)$$

These justify the following approximation of (68):

$$-Z_0(\beta l_{cs} + n_1\pi) = Z_c\beta_c l_c. \quad (71)$$

For the uniform line, the value of l_s is zero and the value of Z_s is Z_0, so that from (53), βl_{ss} becomes $n_2\pi$, and from (71)

$$\beta\Delta S_s = \beta l_{cs} - \beta l_{ss} = -n_1\pi - Z_c\beta_c l_c/Z_0 - n_2\pi. \quad (72)$$

In these measurements, the value of $\beta\Delta S_s$ between zero and $-\pi$ is used, so that

$$\beta\Delta S_s = -Z_c\beta_c l_c/Z_0. \quad (72a)$$

Applying[4] Everitt's (44a) to the terminology of this paper, we obtain, assuming the end capacitance of the line socket equal to that of the fitting under test

$$Z_{co} = -\frac{Z_c[(1/\omega C)\cos\beta_c l_c + jZ_c\sin\beta_c l_c]}{Z_c\cos\beta_c l_c + (1/\omega C)\sin\beta_c l_c} \quad (73)$$

$$X_{co} = Z_c\cot(\beta_c l_c + \tan^{-1}Z_0\omega C). \quad (73a)$$

For the case of an open-circuited fitting, X becomes X_{co} and l_m becomes l_{co}, so that combining (51) and (73a) yields

$$Z_0\tan\beta l_{co} = Z_0/\tan(\pi/2 - \beta l_{co}) = Z_c/\tan(\beta_c l_c + \tan^{-1}Z_c\omega C). \quad (74)$$

From (48), (69), and (70), this is approximately

$$\beta l_{co} \cong n_3\pi + \pi/2 - Z_0(\beta_c l_c + Z_c\omega C)/Z_c. \quad (75)$$

For the uniform line, as shown in obtaining (72), βl_{ss} becomes $n_2\pi$, so that from (65)

$$\beta\Delta S = \beta l_{so} - n_2\pi = \pi/2 - Z_o\omega C. \quad (76)$$

Combining (75) and (76),

$$\beta\Delta S_o = \beta l_{co} - \beta l_{so} = (n_3 - n_2)\pi - Z_o\beta_c l_c/Z_c. \quad (77)$$

Again, the value of $\beta\Delta S_o$ used lies between zero and $-\pi$, so that

$$\beta\Delta S_o = -Z_0\beta_c l_c/Z_c. \quad (77a)$$

Dividing (72a) by (77a), we obtain

$$\Delta S_s/\Delta S_o = Z_c^2/Z_0^2 \quad (78)$$

and since Z_c is always positive,

$$Z_c = Z_0\sqrt{\Delta S_s/\Delta S_o}. \quad (78a)$$

Canadian RTPB

Panel C of the Canadian Radio Technical Planning Board held its first meeting on May 9, 1945, under the chairmanship of Mr. S. Sillitoe. Many of the members of this Panel belong to The Institute of Radio Engineers and are active in its work. Dr. F. S. Howes, the vice-chairman of this Panel, is the immediate past chairman of the Montreal Section of the I.R.E. and is the chairman of the Canadian I.R.E. Council. Mr. S. Sillitoe, the chairman of the Panel, is also a past chairman of the Montreal Section and Mr. L. A. W. East is its present chairman.

CANADIAN RTPB

Front Row—left to right—M. R. Olding (A'45–M'45); L. A. W. East (A'25); J. S. Ford (A'28); J. H. Pratt (A'43–M'45); and C. H. Brereton (A'41–M'45).

Second Row—left to right—J. H. Bolton; R. R. Desaulniers (A'36); J. E. O'Brien; B. McNeil (A'37–M'45); C. W. Boadway; D. B. Black; E. G. Ratz; R. Meadows; and L. Walker.

Third Row—left to right—J. L. Clarke (A'30); F. S. Howes (A'37–M'43–SM'43) vice-chairman; E. S. Kelsey (A'38); A. B. Hunt (A'43–SM'43); E. Olson (A'30); J. Warren (A'40–SM'45); S. Bonneville (A'38); S. Sillitoe (A'35) chairman; and H. S. Finnemore.

In Unity There Is Strength

The I.R.E. and Its Friends

Alfred N. Goldsmith
Editor

There are times in the career of organizations when a mood of sober determination speedily to accomplish the best results may properly give way, for a time, to a mood of satisfaction and jubilation over major achievements. This is such a moment in the life of The Institute of Radio Engineers.

For more than a third of a century the Institute has marched steadily forward, sometimes in difficult days. Like other specialized professional organizations, it was best known to its own membership. The extent of their loyalty was surmised. But it was difficult even to guess the degree to which the radio-and-electronic industry which the members of the Institute serve was itself interested in, and friendly toward the Institute.

Time provided a touchstone to determine otherwise unknown factors and viewpoints. Pressed by the need for adequate headquarters space, the Institute inaugurated its Building-Fund Campaign. It asked from its members, and from the friends of the Institute in industry, a substantial fund. The outcome of its proposal could not be predicted.

The result has been conclusive both in magnitude and in its mode of accomplishment.

The members of the Institute and executives high in industry rallied with enthusiasm and vigor to the furtherance of the Building-Fund Campaign. They gave freely of their time, thought, and energy. A competent office staff collaborated fully with them.

And the Institute's membership responded just as every loyal member of the Institute hoped and expected that it would. Nor were contributions to the fund restricted to the higher grades of membership. Notable showings were made in every grade of membership—and the significance of this in the Student grade, for example, must not be overlooked. It augurs well for any association when its youngest members feel themselves a true part of it and, from limited means, give their quota.

Industry responded in two fashions. The first of these was of financial nature. Substantial contributions from great numbers of organizations, both large and small, in the radio-and-electronic field, built up an encouraging and impressive total. Even more significant was the good will and understanding displayed toward the engineers and their Institute by industry itself. This evoked a similarly appreciative response from the engineers. A long step has been taken toward the integration of the engineers and their industry. Industry is clearly appreciative of its indebtedness to the engineers, and the engineers fully recognize the opportunities of advancement which their industry presents to them.

And so the Institute and its members may properly have a sense of fulfillment and happiness in the recognition which they have gained in a material sense and the high repute which they have achieved.

The glow of satisfaction must be only the introduction to a mood of determination to be fully worthy of the trust, confidence, and co-operative attitude of industry. The engineers in the Institute are alike resolved that their paths shall be those of constructive accomplishment and their services to the public and industry of noteworthy nature. The Institute has successfully ended one epoch in its career. It now enters a new era bright with the prospects of still worthier and more helpful accomplishment.

Alfred N. Goldsmith
Editor

Success: A Challenge to Further Effort

William L. Everitt
President

The Building-Fund campaign has been an overwhelming success. The Institute of Radio Engineers has completed a major project in its forward progress. We are all delighted. We are now assured of adequate housing. But even more important than the physical plant it will provide is the indication it gives of the strength of our membership and the recognition by industry, not only of our past accomplishments but also of our future promise. Success begets success, and gives a stimulation to further effort. We should now move on to new tasks with confidence and assurance. We have also engaged in a critical self-analysis which has shown us where improvements can be made, for you may be sure that during a financial campaign every weakness will be brought to light and pointed out forcefully. Many of these criticisms have been carefully considered and steps taken to eliminate their cause in the future.

During the same period that the campaign has been in progress to furnish the housing, an

adequate headquarters staff has been developed to provide the co-ordination and long-time leadership necessary for future progress. Housing, in keeping with the dignity, prestige, and needs of the Institute, such as the Building Fund will provide, is essential to the proper functioning of this staff and the maintenance of a high esprit de corps.

The holding of membership in a professional society should not resemble a subscription to a magazine, in which money is paid for services rendered and no further obligation is incurred. The Institute is a co-operative endeavor to advance the profession as well as the professional standing of the individual. While the success of the Building Fund will provide more services for the members, it is of paramount importance that the expanded facilities are used as well to provide for the general advancement of the radio, electronic, and allied arts.

The Building-Fund campaign has shown what can be accomplished by the hard work of the members under the able leadership of the various committees and their chairmen. The long-continued success of the Institute in its technical, sectional, and other activities will continue to depend upon the mental and physical effort of the members. The development of adequate headquarters (physical plant and personnel) will put the enormous latent potentiality of our group to more effective use. In successfully carrying through this task we have not earned the right to rest but rather the means for continued service which will call for a wider distribution of tasks among our membership. The Institute of Radio Engineers has an opportunity for service to the industry, to the membership, and to the world. Let us be sure we take advantage of it by the continued application of initiative, intelligence, and hard work.

W. L. Everitt
President

Powel Crosley
Honorary Chairman
Building-Fund Committee

It is with great pleasure and pride that I announce that the Building-Fund campaign of The Institute of Radio Engineers has not only attained, but well exceeded the goal of $500,000 originally established for the purpose of:

1. Expanding the facilities of the Institute.
2. Providing the Institute with a permanent home.
3. Creating a closer union between engineers and their industries.

With comparatively little time in which to raise a fund of this size, such results can be attributed only to the untiring, constant, and sincere efforts on the part of the membership and campaign workers in their earnest determination to make this campaign a huge success.

Lynde P. Wheeler
Honorary Vice-Chairman
Building-Fund Committee

As Honorary Chairman of this very worth-while project, I wish to thank the membership for their endeavors, the sponsors for their support, the campaign workers for their unceasing solications. The Board and management of the Institute have, with a single all-out effort, achieved a double purpose: in having acquired money which, by their own estimate, they needed to undertake the responsibility of a greater Institute; and in having drawn engineers and their industries closer together, to their mutual advantage, than ever before.

Powel Crosley
Honorary Chairman
Building-Fund Committee

The Conception and History of the Building-Fund Campaign

All members of our Institute have cause for jubilation in the momentous success of our campaign to raise funds for a suitable and permanent headquarters building. Since our Institute of Radio Engineers was founded in 1912 our membership has increased almost a thousandfold. The success of our campaign is a gratifying symbol of progress not only for our Institute but also for the radio-and-electronic industry with which we are associated. Ownership of our Institute home will be consistent with the stature and importance of our organization and its growing responsibility to the industry and to society. As we reach this milestone in our history it is appropriate briefly to review the conception, organization, and progress of the campaign now approaching its conclusion.

For a small organization there are economic and perhaps other advantages in leasing headquarters office space. But with continuing growth such advantages may progressively diminish and finally disappear, or be outweighed by other advantages. Professional societies are normally exempted from the payment of real-estate taxes. But they cannot take advantage of such exemption if they lease space in taxable commercial buildings. Therefore, the economic advantages of ownership are greater for a professional society than for a commercial institution, and the advantages of ownership

appear at an earlier stage of its growth.

The Institute has many times had to move its headquarters, change its address, and otherwise adjust itself to the inconveniences of tenant occupancy. In 1943, we had outgrown our present offices, additional space was not available, and we were faced again with the necessity of moving. To find suitable quarters the Board appointed an Office-Quarters Committee, consisting of Mr. Raymond A. Heising, Chairman; Dr. Alfred N. Goldsmith, Dr. William L. Everitt, Mr. Haraden Pratt, Mr. Herbert M. Turner, and Mr. Harold R. Zeamans.

Kaiden-Kazanjian

BENJAMIN E. SHACKELFORD, *Chairman*
Building-Fund Committee

Ray Lee Jackson

RAYMOND F. GUY
Vice-Chairman
Building-Fund Committee

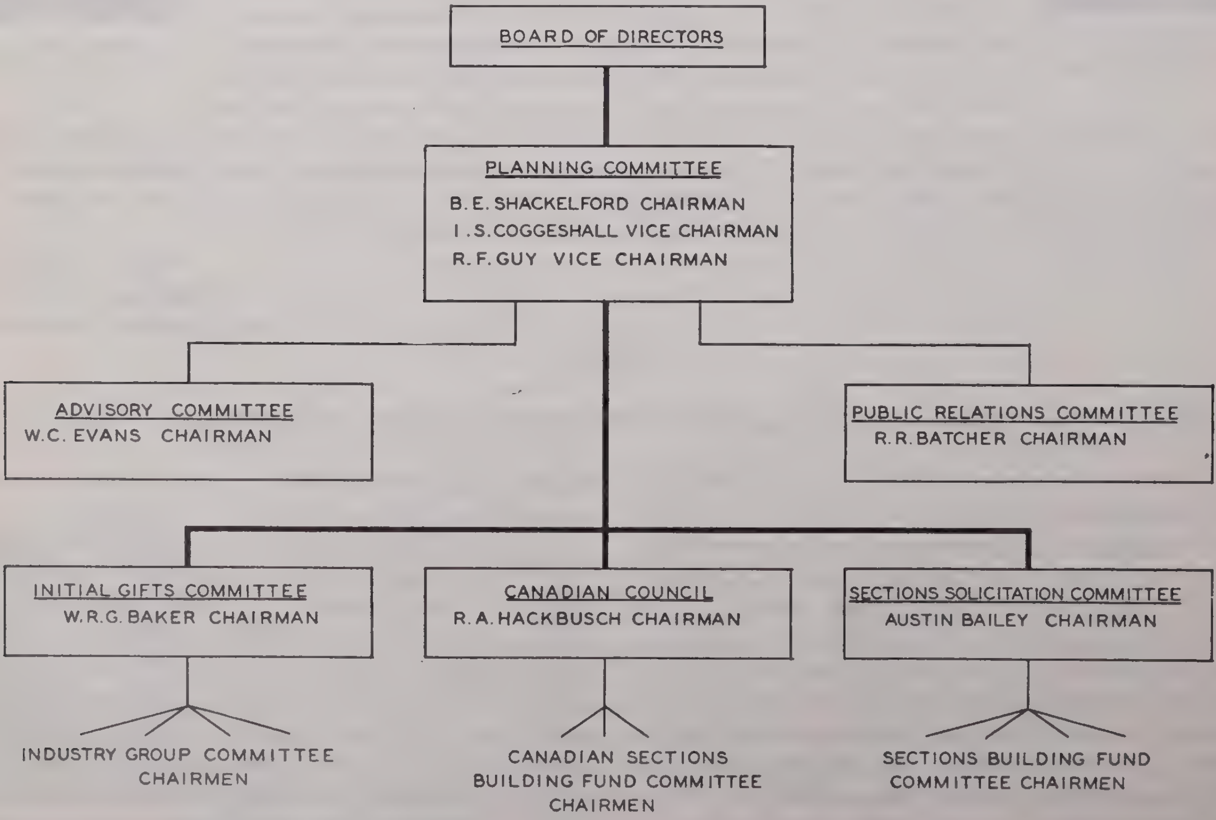
Recognizing that the Institute had attained a stature warranting consideration of the purchase of an adequate headquarters building of its own, this Committee investigated and, in November, 1943, reported to the Board that such a step appeared feasible. After due consideration the Board, at this meeting, gave approval to the general plan for procuring and equipping such a building.

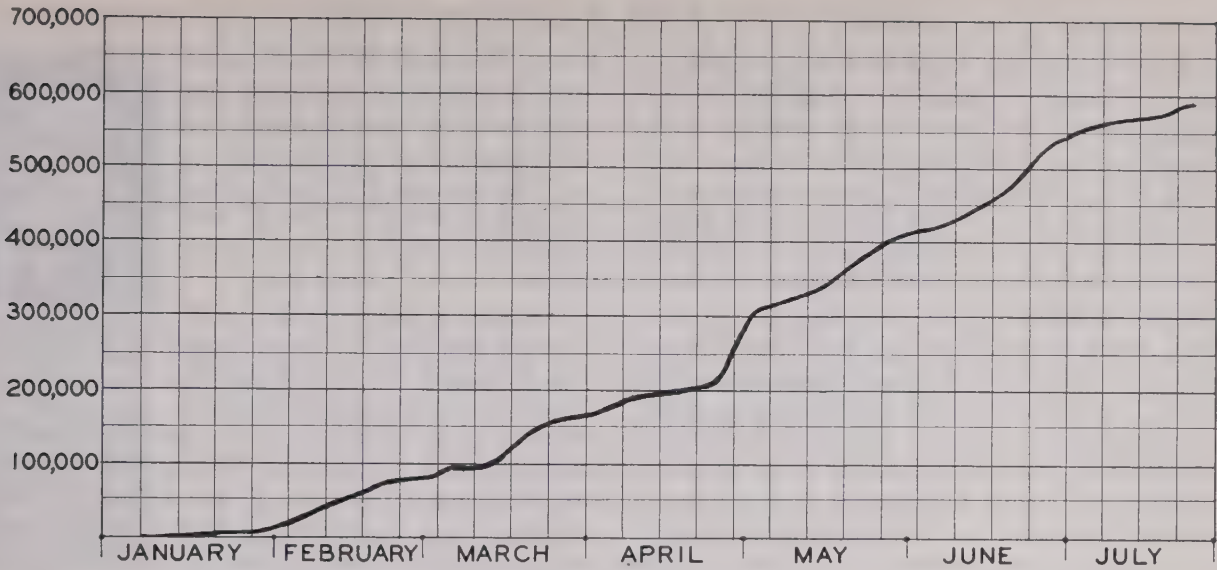
A month later, in December, after further study of the matter, the Board instructed the Office-Quarters Committee to locate a suitable building for purchase and report further on the economics of the project. During the ensuing year the Committee devoted a great deal of time to searches, investigations, and studies and the Board inspected about a score of buildings.

At the January, 1944, Sections Committee meeting, the Board's thinking on the purchase of a building was announced and explained and shortly thereafter the Board created the nucleus of the Building-Fund Committee with Dr. B. E. Shackelford as chairman and Mr. I. S. Coggeshall as vice-chairman. After the inauguration of the campaign, Mr. Raymond F. Guy was also appointed vice-chairman.

IVAN S. COGGESHALL
Vice-Chairman
Building-Fund Committee

The spring and summer of 1944 were devoted to working out details of organization and procedure and accumulating the large amount of information required incidental to the proposed campaign. At the September, 1944, meeting, the Board unanimously approved setting up a Planning Committee, an Advisory Committee, and a Working Committee, approved the minimum goal of $500,000, of which $75,000 was for the membership and $425,000 was for corporate, individual, and other special donors,





GROWTH OF THE I·R·E· BUILDING FUND

and appropriated $500 for preliminary Committee expenses.

The Planning Committee has been the keystone of the organization and has had general direction and charge of all activities. The Advisory Committee advised the Planning Committee on basic policies, concerning public relations and relations with industry. The Working Committee was intended to be a large group divided into many subcommittees concerned with solicitation, publicity, and the like, but in the ultimate evolution of the organization these functions were assigned to the Initial-Gifts Committee, the Sections Solicitation Committee, and the Publicity Committee. The organization of the Building-Fund Committee, as finally constituted, is shown on the condensed organization chart.

The $500,000 goal was arrived at on the basis that somewhat more than half would be available for purchasing, altering, furnishing, and equipping the land and building and the remainder would be invested to give a return which would be applied to maintenance and carrying charges. The Institute paid $4000 annually for rental and it was estimated that the annual maintenance costs for a building would be considerably higher. The difference would be met through the interest on the money invested, and therefore annual operating costs paid out of income from dues would not be increased.

By November, 1944, the Building-Fund Publicity Committee was ready with the publicity material which appeared in the December, 1944, and January and February, 1945, PROCEEDINGS, explaining the need for a headquarters building and announcing the goals of the campaign.

On November 29, 1944, the Building-Fund Committee was authorized to contract with Aderton-Johnson Associates of Harrisburg, Pennsylvania, fund-raising counsellors, for their assistance in planning and carrying out the campaign and $15,000 was appropriated for Committee operating expenses. Building-Fund Headquarters space was rented immediately at 55 West 42nd Street in New York City and preparations to start the campaign were undertaken.

At the January, 1945, meeting, the Board unanimously adopted a resolution defining the Institute's Building-Fund policy, and during the January 24–27 Midwinter Technical Meeting in New York, publicly announced the inauguration of the campaign.

The progress chart shows month by month the growth of the Building Fund. At this writing, it is still growing and the comfortable margin by which the minimum goal is now exceeded will continue for some weeks to become greater. It is of special interest to note that, since the minimum goal has been substantially exceeded, the invested capital to provide building maintenance income may exceed the amount estimated, the return on the investment, correspondingly, may be in excess of the amount estimated, the maintenance charges against dues income may be correspondingly reduced, and the Institute may therefore not only own its own home but pay less for the privilege than it did to rent. Inspection of this chart must inspire a feeling of gratification and accomplishment in this important development in I.R.E. history.

RAYMOND F. GUY
Vice-Chairman
Building-Fund Campaign

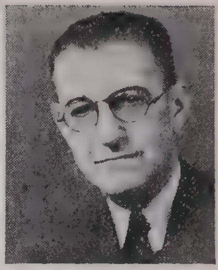
Industry and the Engineer

World War II has had the effect of high-lighting the importance of a rather specialized engineering group—the radio engineers. It offered to this segment of the engineering profession the opportunity to prove their worth not only to their industry, but to their country. History will tell the story of how this engineering group accepted the challenge and performed what will seem, when the facts are available, almost a miracle in engineering accomplishment.

While it is true that radio or "wireless" played an important part, primarily for communication purposes, in World War I, it was insignificant as compared to the radio-and-electronic applications employed in World War II. Unfortunately these new, and in many instances revolutionary equipments and systems, were devised to assist in fighting a brutal war. It is, nevertheless, equally true that out of this work will come new products and and new services for the public and a stronger and broader radio-and-electronic industry.

The executives of the industry recognize and respect the accomplishments of the engineers. The tremendous expansion of the industry so clearly based on new applications and engineering advances precludes any question as to the part played by the engineers of the industry.

It is also most evident that in carrying out this expansion the executives of the industry have demonstrated administrative and organizational skill unsurpassed by any other industry, thus indicating that when peace returns they will be ready to produce the new products and systems which will result from the intensive effort put forth during the war.



W. R. G. Baker
Chairman
Initial-Gifts Committee

The demands made on the industry by World War II, have, therefore, been met through the joint efforts of the executives and the engineers of the industry. As a result there has developed a mutuality of interest which has gone a long way toward deeper integration of the engineers and their industry. This is, indeed, a beneficial and hopeful condition and will assist in easing some of the problems of the industry during the transition from war to peace.

The success of the Building-Fund campaign is tangible evidence that the industry proposes to support its engineers. It further indicates that The Institute of Radio Engineers recognizes its responsibility to the industry and plans to provide facilities for the continuing development of a sound engineering organization.

W. R. G. Baker
Chairman
Initial-Gifts Committee

The Contribution of Members to a Greater I.R.E.

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Austin Bailey
Chairman
Sections-Solicitation Committee

Although the Building-Fund campaign has been carried on by radio engineers and for the benefit of radio engineers, it is, nevertheless, significant that industry has contributed handsomely to assist the Institute in attaining its goal. Industry and the engineer, working together as a team, have done and will continue to do much that is outstanding in this age for electronics. Those things which help the engineers in their profession likewise are of great value to industry and both recognize this interrelationship.

Every member of I.R.E. has been given an opportunity to contribute to this once-in-a-lifetime appeal. Each has responded in a measure commensurate with his individual ability and with his vision of the future of his profession and of the Institute. Each Section has carried through a personal approach to its members and while this is being written new reports of success and added funds are still being received. Section organizations have done a really fine job which will promote and further the aims of the Institute. Members not residing in the territory assigned to sections, men in the Armed Services, engineers from all over the world and student engineers all have been solicited and many have given. The bonds of common interest and the pride of membership in a successful professional organization are the dynamic forces that have brought about our achievement.

As of the beginning of this year the Institute had a total membership of just over 13,000. These were roughly distributed as follows:

In 33 Sections (5 of which are outside the U.S.A.)	8,700
Not residing in Section territory	2,200
Students	2,100
Total	13,000

As of this writing, about 2700 members of I.R.E. have made contributions and more are still coming in. This represents 21 per cent of the entire membership. About 28 per cent of members in Section territory have contributed. In a third of the Sections the number of subscribers was 50 per cent or more than the number of members in those Sections. A number of the Sections exceeded their goals by wide margins. The total of funds

so far received through Sections and from members at large is around $113,000. We should be proud of what we have accomplished and glad to have our names as contributors entered in the roster of those who have had a part in promoting a greater Institute of Radio Engineers. With continued interest and activity we shall now move on to new tasks with assurance that our professional society is recognized as outstanding both by the engineers themselves and by the industry they have helped to promote.

Austin Bailey
Chairman
Sections-Solicitation Committee

Locating a Prospective Home

Raymond A. Heising
Chairman
Office-Quarters Committee

The Office-Quarters Committee was appointed by the Board of Directors at its September 8, 1943, meeting. Its original duties were those of studying the advantages of renting versus home ownership in view of the approaching necessity of securing more space. During the winter of 1943–44 this study showed that the renting situation in New York City was bad. Suitable space would triple our present rent. The advantages of home ownership looked promising in view of low real-estate prices. Several other societies had already purchased, or were planning to purchase permanent homes.

Early in January, 1944, arrangements were made with a real-estate counsellor to locate for us properties that might be purchased. The committee visited and inspected dozens of places, all of which required the expenditure of much time, and the acquisition of dust and dirt from a few of the properties. The limited Institute funds then available for purchase together with estimates of what monies might be collected restricted our choice in buildings, and nothing was found that satisfied the Committee or the Board as a suitable home for the Institute for any extended period of time.

The Committee in its studies looked into the matter of joining the four "Founder Societies" in the Engineering Societies Building. Investigation showed that each of these societies had contributed upward of $265,000 of its own funds. This would presumably necessitate our advancing a like amount if we should join with them. This was likewise beyond our means at the time. It was found that there was some discussion going on among the Founder Societies about establishing an "Engineering Center," erecting a new building, and securing the adherence of more engineering societies that were located elsewhere. Such discussions, however, had not reached the state of formulating plans.

The Committee by this time found itself facing these facts. Some Board members favored the Engineering-Center idea, others favored owning our own building. Either would involve funds greater than our reserves. The Committee could not conscientiously advise committing the Institute to one of these proposals. The former would commit the Institute to an unknown amount (at least $265,000) as plans for an Engineering Center had not been formulated. The latter proposal would require risking some or all of the Institute's reserves on options pending the outcome of a campaign to raise money. Prudence, therefore, dictated that an effort should be made to raise the money before committing ourselves, even though the campaign had to operate under a handicap.

Evidence came to the Committee during this period that now was a propitious time for the Institute to raise money. The Committee, therefore, outlined tentative plans for a fund-raising campaign, and recommended that the Board start such a campaign with the objective of raising a fund "to be used in connection with the establishment of a suitable headquarters building, whether alone or in association with other engineering societies, as the opportunity presents."

From the spring of 1944 until the spring of 1945, the Committee continued its search for a suitable building and kept itself informed on the development of plans for the Engineering Center. By May of 1945, the fund was growing at a rate that would give an indication of its ultimate size and more definite ideas were formulated as to the amount that might be expended for a building. It was found possible to consider some properties that seemed out of the question a year ago.

At the time this is being written, the Committee is taking certain steps to purchase a building under authorization from the Board of Directors. The plans for the Engineering Center have progressed too little to make it advisable to wait, while real-estate prices are definitely on the upgrade. If the Engineering-Center plans move as slowly as such plans usually do, they will hardly take final form within two or three years, and an equivalent time in addition will be required to build a building. The Board, in authorizing purchasing a building, has not dropped the Engineering-Center idea. When plans for the latter materialize to the point that they can be given consideration, the Board will weigh the advantages and the costs against its location at that time.

In the meantime (as of July, 1945), the Committee is following the customary prudent policy of not advertising the properties for which it is conducting negotiations.

R. A. Heising
Chairman
Office-Quarters Committee

A Three-Cornered Conference on The Institute's Future

George W. Bailey
Executive Secretary

Secretary Pratt, Treasurer Heising, and I were sitting on the stairs in the front hall of the building which may become the new home of the Institute. We had just completed a tour of the building, from cellar to attic.

"Well, George," said Secretary Pratt, "can you visualize the rooms as they will look when all the personnel are at their desks?"

"Yes," I said, "I certainly can, and I can also imagine the satisfaction of our members."

"I am sure the members will like such a building as this," said Treasurer Heising. "It is handsome on the outside, and these big rooms on the first floor will look inviting when they are furnished for the convenience and comfort of visiting members."

"That is true," I said. "For the first time, members can visit headquarters and not feel they are in the way."

"Ever since I can remember," said Secretary Pratt, "I always felt that I should not go to the Institute's office unless it was really necessary. There never had been a place to sit and read or write, make a phone call, or hold a conversation with another visiting member without interfering with an office worker, borrowing somebody's chair or telephone, or feeling that your presence was annoying to others. Also, when a committee wanted to meet, it was necessary to bother someone and interfere with his work in order to create space in which to hold the meeting."

"Now," I said, "members can come freely and feel that they are welcome because it will be possible for them to read, write, telephone, confer, and hold committee meetings without interfering in the least with the functions of any of the office staff, which is as it should be. Our members are entitled to such facilities. And think how convenient it will be for those coming from a distance, from abroad, for instance. However, speaking of the advantages to members, I meant more than just these things. I was referring to the fact that our new home will enable us to conduct our affairs so that the members will directly benefit by them."

"What are some of those advantages?" asked Treasurer Heising.

"There are so many," said I, "it is difficult to enumerate them in the order of their importance, so I'll just mention a few as they occur to me. First of all, there is the fact that the Editorial and Secretarial offices will now both be in the same building instead of sixteen blocks apart as at present."

"That's important," said Secretary Pratt, "and I expect Dr. Goldsmith will tell us what it means to his Editorial Department to have room enough to carry out their projects, and the advantage of accessibility to the personnel and records of the Secretarial Department."

Haraden Pratt
Secretary

"Dr. Crew, our Assistant Secretary," I said, "will be able to get a good start in his newly acquired duties because he will have adequate space and the tools with which to function."

"I don't see how you found a place for him in the old office," Secretary Pratt said.

"Well," said I, "he was allotted the last few square feet, and, as it was, he scarcely had room to turn around. And his secretary, Miss Graham, will be glad to have plenty of desk space."

"We are receiving many letters from members in the Service who are planning their postwar education and wish advice. Dr. Crew and I are prepared to answer such letters, and welcome inquiries from our members."

"We are also planning to give advice and assistance to members seeking postwar jobs where their qualifications can be used to best advantage. At the same time, we shall be prepared to assist industrial concerns and academic institutions in locating scientific and technical personnel. One of our long-range projects is the maintenance of an active and reliable file of electronic engineers and technicians."

"Now, I believe you and Dr. Crew can undertake the project of assisting Sections in obtaining speakers," said Treasurer Heising, speaking in his capacity as Chairman of the Sections Committee, "and, also, you will have a chance to work out plans for a traveling lectureship to go into effect when transportation difficulties are overcome."

"That's right," said I, "and Dr. Crew will be helpful in office affairs also, such as working out a group insurance and pension plan for the office staff."

"Good," said Secretary Pratt. "A permanent home for the Institute is going to be an incentive to our office staff to make plans for a career with us. Several of the staff have been with us a long time, and I hope the others will continue with us. New office quarters are extremely useful, but, after all, what counts most is the experience and skill acquired by long-term office workers."

"You know we have been planning to have a Technical Secretary," said Treasurer Heising. "I don't quite understand why we are so slow in selecting him."

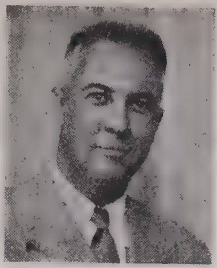
"Because," I said, "we had no place to put him."

"Now we can go ahead," observed Secretary Pratt, "and that will enable us to achieve several objectives of advantage to our members and the enhancement of the prestige of the Institute."

"How do you figure that?" said Treasurer Heising.

"The Technical Secretary," replied Secretary Pratt, "will be able to devote a large amount of time to the

work of the technical committees and will be of great assistance to them. He is to have supervision of standardization activities and will help to organize the right committees with appropriate persons serving on them. Fortunately, Mrs. Fisher, who has been with us some time, is ready to act as his secretary, so that problem is solved."



William H. Crew
Assistant Secretary

"One of these rooms will make a wonderful library and reading room," said I. "Visitors will not have to ransack the place to find a back number of the Proceedings."

Alice Connolly
Chief Accountant

"What about the conduct of the affairs of the office itself," said Treasurer Heising, "How will it be improved?"

"Miss Lehmann, our Office Manager," I answered, "at last will have the opportunity to carry out many of her ideas of office arrangement and routing, which were impossible in the old office due to crowded quarters. Take, for instance, the Addressograph Department, an important part of a membership society like ours. With many new members every month, and a large number of changes of address, particularly of those in the Service, accuracy and promptness are essential. We now have a marvelous, big, new, fast machine. It is a mystery to me how Miss Reinhardt and her assistant, Miss Boersma, ever manage to run it in the little cubbyhole where they are now located. They will surely make good use of that big room we have allotted to them."

"Let's see," said Treasurer Heising, "right across the hall from that room will be the mail room, is that it?"

"That's it," said I, "right on the ground floor, near the side entrance. And Joe Saitta and Miss Clemens will be mighty glad to have room to sort mail. Joe will have adequate space for that big automatic parcel-post weighing machine and the postage-meter mailing machine. Now he can turn the crank of the duplicating machine without bumping his elbow on a steel filing cabinet."

"I expect you can make much more prompt shipments of back numbers of the Proceedings and supplies," said Secretary Pratt, "now that the store room is next door to the mail room and not eighteen floors away as it was before."

"You are right," I answered, "and, speaking of being prompt, that really will apply to all our correspondence. We are acknowledging all current letters right away, but the real test is going to be the speed with which we can furnish information requests. That is where Miss Lehmann's plans on filing will come in."

"You know, the accessibility of facts is the key to prompt answers to correspondence. With our new arrangement of files, Mrs. Burke and Miss Finnigan, with their temporary assistants, will have an up-to-date filing system."

"I am sure," commented Secretary Pratt, "that Miss Lehmann will be glad to have her new office with her secretary, Miss Woolbright, where she will be near her correspondents, Mrs. Dalton and Mrs. Harding."

"Yes," said I, "and also Miss Roosevelt, who has charge of applications, with Miss Pohorely, her assistant; and Mrs. Sablesak, in charge of readmissions, will all be glad to get rid of their present severe handicap of cramped quarters. And there is Miss Neumayer, our statistician, and Miss Johnson, who handled subscriptions and stenographic work for the Accounting Department."

Elizabeth Lehmann
Office Manager

"I expect," said Treasurer Heising, "that Miss Connolly, our Chief Accountant, is going to find a great difference in carrying on her duties in her new office."

"I should say so," I said, "and her staff, Miss Tecchio, Mrs. Stone, Miss Gallagher, Miss Desposito, and Mrs. Giles are all going to make good use of more space. And they won't all have to stop work to let Auditor Todt into the office!"

"I hope and expect that the Telephone Company will be able to supply us with a larger board for the new building without too much delay, so that Miss Andreyo will be able to handle more calls promptly."

"The Directors," said Secretary Pratt, "will be pleased with that attractive room where they can hold their meetings and have all the office records easily accessible. Besides, this room will be very handy for large committee meetings. We have never been able to accommodate our larger committees adequately before."

"That's a fact," said I, "and I am sure the officers will like the executives' room that is planned. The next room to it can be used for small committee meetings, and Mr. Copp will find it convenient to interview his advertising clients there."

"How about your office?" asked Treasurer Heising.

"Well," I replied, "that corner room on the second floor would be very suitable for me and my competent secretary, Mrs. Radigan. I certainly am lucky in having a secretary who knows the history and office routine of the Institute so well.

"It must be a tremendous satisfaction to you two officers, who, with the others, have worked so long for the Institute, to see your dreams come true."

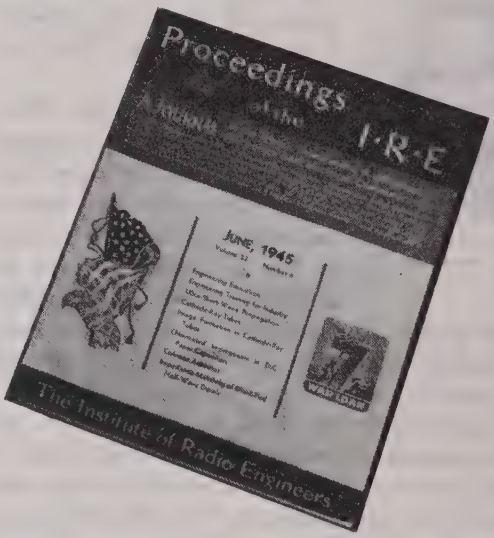
"Yes, it is," they both agreed, "thanks to every one of the donors who made it possible. But, now that it has been achieved, we must not forget that it is the responsibility of all of us to put our shoulders to the tasks that lie ahead, and make our Institute the bigger and better organization that the phenomenal growth of radio and its allied fields of engineering demands and requires."

George W. Bailey
Executive Secretary

The Editorial Department of the Institute feels obliged to present the following communication from the PROCEEDINGS, in the form in which it was received.

THE PROCEEDINGS SPEAKS FOR ITSELF

"How do you do? I am the PROCEEDINGS. I've just arrived on your desk and I'm sure you'll agree that I'm full of informative material on the latest developments in radio and electronics. Please read me carefully, all of me, the technical and nontechnical articles and the advertising. (Mr. Will C. Copp and his staff really worked very hard to present the best possible advertising material to you; and Dr. Goldsmith and my good friends in the Editorial Department took great pride in seeing that my pages were full of papers which would prove of greatest value to you and engineering material which they and I are sure you will find interesting.)



"Where was I born? Who brought me up? And how did I reach you? Well, Mr. Dorman Israel and his Vice-Chairman, Mr. E. T. Dickey gathered together a group of sixteen subchairmen called the Papers Procurement Committee, who lie awake nights wondering where they can go to get the best possible papers on the subjects in their special fields. Then too, some authors, of their own helpful volition, send us manuscripts, and again, someone writes in that he knows of a good paper.

HELEN M. STOTE
Associate Editor

"So, after a time, the manuscript of each paper is sent in to the Editorial Department where Mr. R. D. Rettenmeyer, my new Technical Editor, classifies the paper and makes out elaborate cards and forms concerning me —I really have no privacy at all; and then he sends me to Dr. Alfred N. Goldsmith. (If Dr. Goldsmith thinks that I am in the least dangerous to the war effort, he quietly sends me to a referee, or to the Military Services, for 'clearance.' If I am not cleared, then back I go to the author.)

"But if I have a clean bill of health, I am entrusted again to Mr. Rettenmeyer. He sends me out to at least three different readers of the Papers Committee, who read me and return me with their amazingly searching and helpful comments. Then out I go again to a member of the Board of Editors, together with all the comments of the Papers Committee readers. The member of the Board of Editors looks me over and reports his opinion of me. Then back I go to Dr. Goldsmith again with everyone's remarks. By this time I feel as though I were living in a goldfish bowl. Dr. Goldsmith, who is the Editor, then decides whether I should be accepted—good work! sent back for revision—here's hoping! or, more rarely, rejected—hard luck! Fortunately I have many friends who are most capable engineers and authors.

R. D. RETTENMEYER
Technical Editor

"If I am accepted, Miss Winifred Carrière takes me over and writes all over my nice clean pages. She puts on all sorts of mysterious signs for the printer, corrects my English, fixes up my footnotes, arranges for cuts to be made, and in general, prepares me to go to the printer. (As an aside, I might add that sometimes her remarks are not very complimentary, particularly when an occasional author is a bit careless or incomplete, or equations are particularly long and involved, or when drawings are bad.)

Bachrach
WINIFRED CARRIÈRE
Assistant Editor

"After she has done her worst, she sends me to the printer and he sets me in type. Then the printer sends two galley proofs to the author, one to the National Bureau of Standards for decimal classification, one to Dr. Goldsmith, and one to the Editorial Department. Everyone of these people, except the Bureau, reads me, and puts more remarks on my pages, correcting my proof. Eventually, along with the other manuscripts, who have suffered just as I have, I go back to the printer to be put into page-proof form. At this point Miss Helen Stote takes hold of me and adds a lot of material to the galley proofs—front cover, contents page, Sections meetings, Institute Notes, contributors' biographies, and so on. She makes me feel a bit like the filling in a sandwich, for I am in the middle of all this.

"Eventually, page proofs come back; and again Miss Stote does many things to me. She checks over all the galley proofs and compares them against the page proofs, reads and corrects all the material in the Institute Notes section, and dummies that again if it needs it. By the time she gets through with me, I really am exhausted, for she searches high and low for errors and finds most of them. (Occasionally, though, I manage to get the better of all my editorial readers, and slip over an error or two. Even PROCEEDINGS are human!)

WILLIAM C. COPP
Advertising Manager

"During all of this goings-on, Mrs. Ruth Winters toils at her typewriter and writes innumerable letters to authors, the printer, and no end of other people. Mrs. Eleanor Caldwell also gets a crack at me, for she is pressed into service when proofs come in and cuts me to pieces and pastes me up and does some of the proofreading. However, she is pretty considerate as far as I am concerned, since she is mostly concerned with YEARBOOK work and has to leave me in peace most of the time.

"Of course, there are lots of oddments connected with my manufacture which I shan't bother telling you about now. Some of them are appeals to the War Production Board, trying to get very necessary paper for me. I have a healthy appetite for paper, and WPB is certainly strict on points. So Dr. Goldsmith writes volumes of persuasive pleas to the WPB—and sometimes we get some paper. Everyone in the Department is on the 'phone a great deal, checking up on proofs, cuts, the printer, and lots of other people. Reams of letters go out on one subject or another; reports go to different people—there is never a dull moment.

"While all of this has been going on, Mr. Copp has been busily engaged in obtaining advertisements for me. I really feel rather impressive with lots of red, blue, yellow, and green ink in my advertising pages. In addition to being particularly colorful, my advertising pages are extremely helpful to the technical reader. 'Handsome is as handsome does' most certainly applies to me.

"Over 200 advertisers have told you their stories in the PROCEEDINGS and the Winter Technical Meeting program. These advertisements were obtained by Mr. Copp and his associates. A staff of seven in New York, three field representatives in Chicago, Los Angeles, and San Francisco do the service work, prepare market statistics, and handle production detail associated with this major job. Miss Lillian Petranek, Herbert White, and Everett L. Hart assist Mr. Copp with the Eastern selling, while Scott Kingwill handles the Central states, and Duncan A. Scott and Forrest Pearson deal with the growing Pacific Coast firms.

"A new development is an Industrial Research Division in which Miss Eleanor Durkee collects and organizes all kinds of facts about the firms and products serving radio and electronics. Much information of value to the Institute executives, staff, and membership will result from these careful studies.

LILLIAN PETRANEK
Assistant Advertising Manager

"The staff is completed by Leonard Garbin on printing production and Mrs. Julia Winterfield, correspondence secretary. I found that Mr. Copp and his staff have as their guiding principle: Advertisements should present useful facts; engineers will thrive on them.

"So now that you know all about my home training, I hope you will read me with even greater interest, and find my pages more instructive and stimulating than you ever have before. If you have any criticism or complaint to make, send it in. My friends in the Editorial Department will do their best about it. And if you feel like an occasional compliment or pat on the back to the authors, to me, or to the Department, they and I will certainly like it. I look forward to seeing you again soon. Happy reading!"

NOTE: The copy for this section of the PROCEEDINGS, signalizing the success of the Building-Fund campaign and forecasting a greater future for the Institute and the industry, was prepared in July, 1945, before the end of the war with Japan.

THE PROCEEDINGS

Thomas Coke Knight

I.R.E. OFFICE STAFF

Left to Right, Top Row: F. Gallagher, *Order Clerk;* E. Harding, *Correspondent;* M. Finnigan, *File Clerk;* T. Burke, *File Supervisor;* J. Saitta, *Mail Clerk and Office Boy;* Z. Grant, *File Clerk;* D. Pohorely, *Application Clerk;* H. Boersma, *Addressograph Clerk;* L. Reinhardt, *Addressograph Supervisor;* V. Bonn, *File Clerk.*

Middle Row: E. Clemens, *Junior Addressograph Clerk;* F. Giles, *Subscription Clerk;* H. Dalton, *Correspondent;* D. Radigan, *Secretary to Mr. Bailey;* E. Lehmann, *Office Manager;* A. Connolly, *Chief Accountant;* E. Stone, *Assistant Bookkeeper;* M. Lanni, *File Clerk;* M. Andreyo, *Receptionist.*

Front Row: H. Roosevelt, *Application Department Supervisor;* H. Graham, *Secretary to Dr. Crew;* S. Sablesak, *Correspondent;* E. Fisher, *Section Correspondence Clerk;* J. Desposito, *Membership Ledger Clerk;* J. Tecchio, *Cash Disbursement Clerk.*

Not in Picture: A. Johnson, *Correspondent;* E. Neumayer, *Statistician;* M. Woolbright, *Secretary to Miss Lehmann.*

I.R.E. People

W. G. H. Finch

W. G. H. Finch

The United States Navy disclosed on July 6, 1945, that it has promoted Commander W. G. H. Finch (J'16–A'18–M'25–SM'43) to the rank of Captain. Captain Finch is on the staff of Commodore J. B. Dow, head of the electronic division of the Bureau of Ships, United States Navy. Captain Finch was assistant chief engineer of the Federal Communications Commission in 1934 and 1935. He is founder and was the president of Finch Telecommunications, Inc., at the time he went into active duty prior to Pearl Harbor, 1941. Since then he has been on a number of foreign missions to the European theater.

T. A. M. Craven

It was announced on June 14, 1945, that T. A. M. Craven (F'29) of station WOL, has been elected a director-at-large of the National Association of Broadcasters for medium stations. Commander Craven received a majority of the votes cast on a first ballot of a referendum election.

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C. Frank Miller

C. Frank Miller (A'41), formerly a member of the electrical engineering staff of Johns Hopkins University, has been appointed chief engineer of Price Brothers Company, of Frederick, Maryland. Dr. Miller received his Bachelor's degree from Johns Hopkins in 1935 and the degree of Doctor of Engineering from the same university in 1942.

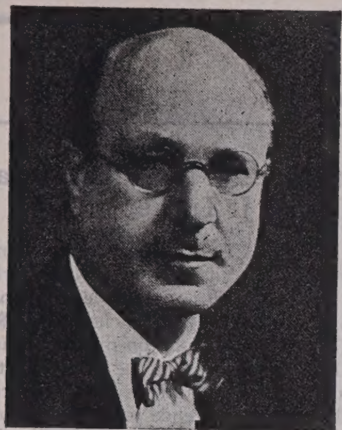
He has been active in the electrical field and in addition to conducting classes at the university, undertook extensive research and experimentation work during the present war emergency in connection with electronic applications involving new industrial processes for organizations throughout the country.

Dr. Miller is a member of the American Physical Society, American Institute of Electrical Engineers, Sigma Xi, and the American Rocket Society.

C. Frank Miller

T. A. M. Craven

I.R.E. People



Daniel E. Noble

Daniel E. Noble

The promotion is announced of Daniel E. Noble (A'25–SM'44) to the position of general manager of the communications and electronics division of Motorola Radio. In his new position Mr. Noble will have direct authority over the engineering, sales, and engineering production departments of the division. He will retain his present responsibilities as director of research.

Mr. Noble's radio experience dates back to his amateur radio station activity before the first World War. He continued work in the radio field through his engineering training at Connecticut State College and Massachusetts Institute of Technology. While teaching in the School of Engineering of the University of Connecticut he carried on extensive radio work as a consultant. His experience with frequency-modulation methods has been continuous since 1936. In 1937–1938 he developed a 100-megacycle frequency-modulation broadcast system, which was used to relay programs from Storrs to Hartford, Conn., for rebroadcast on WDRC and WTIC.

During 1937–39 Mr. Noble developed Connecticut's first frequency-modulation commercial broadcast system with F. M. Doolittle, of WDRC. He also developed the first frequency-modulation police system of this type and the first state-wide police two-way system for Connecticut, completing it in 1940.

Since 1940, Mr. Noble has been director of research for the Galvin Manufacturing Corporation, of Chicago, manufacturers of Motorola equipment. He directed the development of their standard line of frequency modulation radio communications equipment, and the SCR-300 frequency-modulation walkie talkie. During the past two years his division has been concerned with extensive development of radar apparatus.

As chairman of Panel 13, of the Radio Technical Planning Board, Mr. Noble guided the committee work and the preparation of the allocation hearing material for the mobile emergency services and for the many new mobile services proposed. He is one of the members of the FCC-Radio Industry engineering committee appointed to study the frequency-modulation broadcast allocation propagation problem.

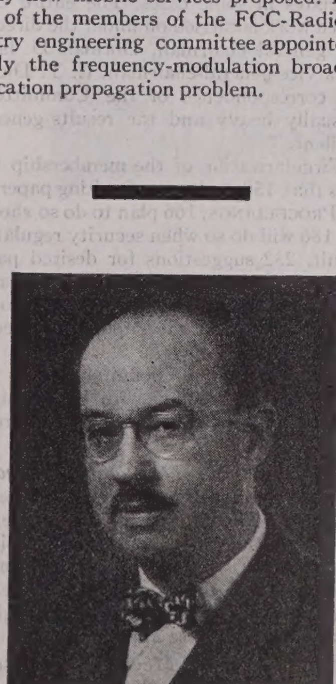
Roy Thurlby Griffith

Roy Thurlby Griffith (A'24–SM'45), assistant engineer of transmission, Western Area, The Bell Telephone Company of Pennsylvania, died at his home in Mt. Lebanon, Pa., on June 23, 1945. He was born at Chester, Pa., on February 26, 1897, and received his education in the public schools and at Haverford College.

He had been employed by The Bell Telephone Company of Pennsylvania since June, 1918 and had been assistant engineer since January, 1927.

As assistant engineer of transmission, he was responsible for the general technical design of telephone facilities throughout Western Pennsylvania to insure satisfactory communication from any telephone to another telephone, including connections with the nationwide network of the Bell System.

Mr. Griffith was a charter member of the Pittsburgh section, Institute of Radio Engineers, and served as chairman of the Section in 1932–33. During the past six months he



Roy Thurlby Griffith

had presented a paper entitled "Telephone Service for Vehicles and Vessels i- Urhar Areas" before the I.P

Clark Named NBC Video Operations Supervisor

Robert W. Clark (A'39), station engineer at the WEAF transmitter, has been appointed television operations supervisor effective June 16, 1945, it has been announced by O. B. Hanson (A'18–M'27–F'41), NBC vice-president and chief engineer.

Mr. Clark, who will report to Robert E. Shelby (A'29–M'36–SM'43), NBC development engineer, will be responsible for technical phases of field and studio operations.

Mr. Clark was born in California, and received the A.B. degree in 1927, and the E.E. degree in 1928, from Stanford University. He joined R.C.A. Communications Inc., in 1928, and was transferred to the San Francisco office of NBC in 1931. From 1932 to 1937 he was assistant station engineer of the KPO transmitter station. He was brought to New York in 1937 to study television engineering, and contributed a great deal to the development and modification television equipment now in use by NBC. He worked actively in major war projects during 1942 and 1943, and in the latter year was named station engineer of the WEAF transmitter.

Correction

On page 415 of the June, 1945, issue of the Proceedings, the membership grades for Mr. Harold H. Buttner were incorrectly listed. They should be as follows: M'27–SM'43–F'44. We take this opportunity of extending to Mr. Buttner our apologies for having omitted the notation of the Fellow grade. *Editorial Department*

Donald W. Short

Donald W. Short (A'27–M'37–SM'43), communication and industrial equipment endineering department, RCA Victor Division of Radio Corporation of America, Camden, New Jersey died on June 5, 1945. He was born in Dubuque, Iowa on January 4, 1906, and attended schools in Lincoln and Dubuque. In 1928, he joined the General Electric Company, Schenectady, N. Y. as a student engineer, at which time he also took courses in general engineering and in radio theory at both the General Electric Company and at Union College. In 1931, he joined the Jenkins Television Corporation, which was later absorbed by deForest. In 1934, after a year with Hygrade Sylvania, he became a design engineer in aviation radio at the RCA Victor Division of Radio Corpo-ration of America, Camd past several year

Institute News and Radio Notes

Executive Committee

July 11 Meeting: The Executive Committee meeting, held on July 11, 1945, was attended by W. L. Everitt, president; G. W. Bailey, executive secretary; S. L. Bailey, W. H. Crew, assistant secretary; Alfred N. Goldsmith, editor; R. A. Heising, treasurer; and Haraden Pratt, secretary.

Membership: Nineteen applications for transfer to Senior Member grade; eight for admission to Senior Member grade; thirty-two applications for transfer to Member grade, thirty-one for admission to Member grade; one hundred and forty-one applications to Associate grade, and eighty-nine applications to Student grade were approved and will be found listed on page 38A of the August, 1945, issue of the PROCEEDINGS.

Admissions Committee: It was unanimously approved that the following suggestions be entered in the Minutes of this meeting, and a draft be sent to the Admissions Committee with an invitation to the Chairman to attend the August Executive Committee meeting: "The Executive Committee proposes the following guiding considerations to the Admissions Committee in relation to elections to the grades of Member and Senior Member: (a) The field of interest of the Institute and its publications, as approved by the Board of Directors, includes the theory, practice, and applications of electronics and electrical communications involving also, among other topics, radio- and audio-frequency measurements, sound and picture electrical recording and reproduction, power and manufacturing applications of radio-and-electronic technique, and industrial electronic control and processes. (b) The interpretation of the term 'allied fields' should be consistent with the Board ruling of paragraph (a) above. (c) The following qualifications should be favorably weighted in considering the membership qualifications of the candidates: past activity in the radio-and-electronic field as indicated by individual contributions or continued work by the candidates, or both; current interest in, and scope of, activities of the candidates in the radio-and-electronic field; present executive standing of the candidates, and control and supervision by them of work in the radio-and-electronic field. (d) For favorable action, it should not be requisite that the candidates shall be engaged solely or even primarily in radio-and-electronic matters provided in general that his other ...ivities are of professional character."

... Seymour Cohn was ap-
...mittee. The fol-
... ap-

ELECTROACOUSTICS

B. D. Bower	R. A. Miller
S. J. Begun	G. M. Nixon
R. P. Glover	Benjamin Olney
F. V. Hunt	H. F. Olsen
V. N. James	H. H. Scott

ELECTRONICS

D. E. Marshall

FACSIMILE

J. C. Barnes	C. N. Gillespie
F. R. Brick, Jr.	J. V. L. Hogan
Henry Burkhard	Hugh Ressler
J. J. Callahan	Arthur Rustad
A. G. Cooley	Lt. L. G. Stewart
R. C. Curtiss	W. E. Stewart

R. J. Wise

TELEVISION

P. J. Larsen	D. L. Jaffe

Proceedings Matters

The Papers Procurement Committee has been fully organized. It consists of 16 groups, and 15 subgroups dealing with specific divisions of the radio-and-electronic field. Each group and subgroup has its chairman, with associates in some cases. The entire Committee work is carried on under the direction of a general chairman, Dorman D. Israel, and a vice-general-chairman, E. T. Dickey. The correspondence of the Committee is unusually heavy and the results generally excellent.

Circularization of the membership indicates that 150 authors are writing papers for the PROCEEDINGS; 166 plan to do so shortly; and 186 will do so when security regulations permit. 232 suggestions for desired papers have been received. All of these data are being sent for appropriate action to the group chairmen of the Papers Procurement Committee.

The membership definitely views with interest papers dealing with the welfare of engineers. These will accordingly be stressed so far as possible.

New Reference Forms and Procedure: It was unanimously approved that the new reference forms be used for Senior Member and Member grade applications, copies of which forms, together with the transmittal envelopes, had been distributed. The procedure for handling such forms through the Sections was likewise approved.

1946 Technical Meeting: Because of prevailing transportation conditions, the dates for the 1946 Technical Meeting have been tentatively postponed to June 26, 27, ... and 29, 1946.

Books

Introduction to Practical Radio by D. J. Tucker

Published (1945) by The Macmillan Company, 60 Fifth Avenue, New York 11, N. Y. 317 pages+4-page index+xvi pages. 155 illustrations. $8\frac{1}{2}\times5\frac{3}{4}$ inches. Price, $3.00.

This is not a book on radio but, as the author makes clear in the preface, a treatment of fundamental electrical principles, with especial reference to the needs of the radio experimenter. Pictures of radio-circuit elements and electrical-measuring instruments appear in the text and reference to radio circuits is made in the problems, but the purpose of the book is held firmly in mind, namely, to provide in a single text, a treatment of the basic principles necessary for an understanding of the more advanced and specialized study of radio applications and circuits.

The subject matter of the book covers not only material usual in textbooks on elements of electricity and magnetism, but chapters are also devoted to the fundamentals of alternating currents and alternating-current circuits. Mathematical digressions into the fields of algebra, simple theory of logarithms, and elementary trigonometry are made at various points in the text where a proper understanding of the electrical theory rests upon a certain minimum of mathematical knowledge.

The style of presentation is clear and simple and, with the mathematical aids provided, should be readily understood by the general reader. Numerical examples are worked out and review questions and exercises, together with the answers to the problems, accompany each chapter. A collection of useful functions and logarithmic tables are provided.

To cover adequately such a wide range of subjects in a bare three hundred pages is no easy matter. It is inevitable that certain omissions and superficialities may be noted. For example, the distinction between power and energy is not always clearly made, and the definition of magnetomotiveforce lacks clearness. It is unfortunate that the usual definition of direction of current flow has been abandoned for the direction of actual electron flow. On the whole, however, the author has produced a comprehensive, interesting and readable textbook, which should be useful for supplying a basis for more advanced study.

FREDERICK W. GROVER
Union College
Schenectady, N. Y.

Contributors

R. R. Bush

R. R. Bush (A'43) was born in Albion, Michigan, on July 20, 1920. He received the B.S. degree in electrical engineering from Michigan State College in 1942. Since that time, he has been a research engineer, working on ultra-high-frequency electronics, at RCA Laboratories, Princeton, New Jersey. He is a member of Sigma Xi, Tau Beta Pi, Phi Kappa Phi, Pi Mu Epsilon, and the American Physical Society.

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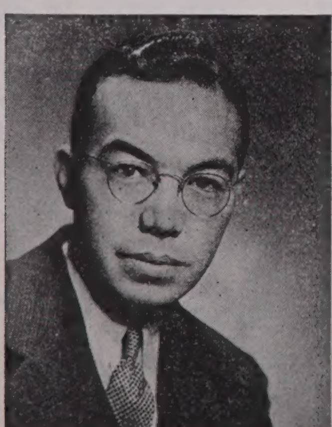
Howard A. Chinn (A'27–M'36–A'42–SM'45) was born in New York City on January 5, 1906. He attended the Polytechnic Institute of Brooklyn, later going to the Massachusetts Institute of Technology where he received the S.B. and S.M. degrees in 1927 and 1929, respectively. From 1927 to 1932 he was a research associate at the Massachusetts Institute of Technology. Mr. Chinn became associated with the Columbia Broadcasting System in 1932 as a radio engineer; from 1934 to 1936 he was assistant to the director of engineering; from 1936 to date he has been chief audio engineer, although, during the war years, the bulk of his time has been devoted to other activities associated with the war effort. From the beginning of 1942 to the end of 1943 he was technical co-ordinator of the Radio Research Laboratory of Harvard University, at Cambridge, Massachusetts, which is sponsored by the Office of Scientific Research and Development. From 1944 to date, he has been first a technical aide and currently a consultant to Division 15 of the Office of Scientific Research and Development. From 1939 to 1941, Mr. Chinn was a special lecturer in electrical engineering at the graduate school of New York University.



Howard A. Chinn

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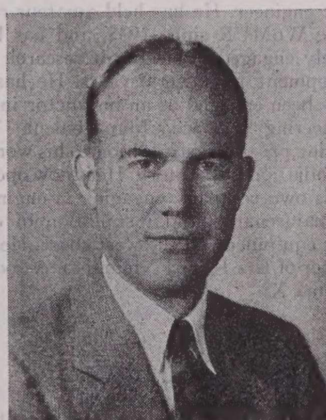

Murray G. Crosby

Murray G. Crosby (A'25–M'38–SM'43–F'43), was born at Elroy, Wisconsin, on September 17, 1903. He attended the University of Wisconsin from 1921 to 1925, receiving the B.S. degree in electrical engineering in 1927 and his professional electrical engineering degree in 1943. From 1925 to 1944 he was research engineer for the Radio Corporation of America in the communications division of RCA Laboratories, where he specialized in frequency modulation. In 1944 he left that position to take up a practice of consulting engineering. He joined the Paul Godley Company, consulting radio engineers, in 1945.

In 1943 and 1944 he served as expert technical consultant to the Secretary of War, and received official commendation for his work.

Mr. Crosby was a recipient of the Modern Pioneers Joint Award from the National Association of Manufacturers in 1940. He is a Fellow of the Radio Club of America and a Member of the American Institute of Electrical Engineers. He is a member of the 1945 I.R.E. Papers Committee, the Papers Procurement Committee, the Admissions Committee, and the Technical Committee on Frequency Modulation.

Philip Eisenberg

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Philip Eisenberg was born on October 2, 1912, in New York City. He received the B.A. degree from the College of the City of New York in 1934. Columbia University awarded him the M.A. degree in 1935 and the Ph.D. in psychology in 1937. He is presently engaged as a research psychologist at the Columbia Broadcasting System, Inc. Previously, he taught psychology at Brooklyn College, undertook a research program in

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W. R. Ferris

E. W. Herold

achievement and intelligence tests for the New York City Board of Education, and installed and developed employee-selection aptitude tests for the War Manpower Commission, in Pennsylvania. He is a member of the American Psychological Association and Sigma Xi.

W. R. Ferris was born on May 14, 1904, in Vigo County, Indiana. He received the B.S. degree in electrical engineering from Rose Polytechnic Institute in 1927, and the M.S. degree in electrical engineering from Union College in 1934. He also did graduate work at Polytechnic Institute of Brooklyn.

He was employed in the research laboratories of the General Electrical Company from 1927 to 1930, when he became associated with the RCA Manufacturing Company in Harrison, New Jersey, as a vacuum-tube engineer. Mr. Ferris transferred from this position in 1942, to the RCA Laboratories, in Princeton, New Jersey, where he has remained to date.

He is a member of Sigma Xi and the Franklin Institute.

Rodney W. Johnson

E. W. Herold (A'30–M'38–SM'43) was born on October 15, 1907, in New York City. He received the B.S. degree from the University of Virginia in 1930, and the M.S. degree from Polytechnic Institute of Brooklyn in 1942.

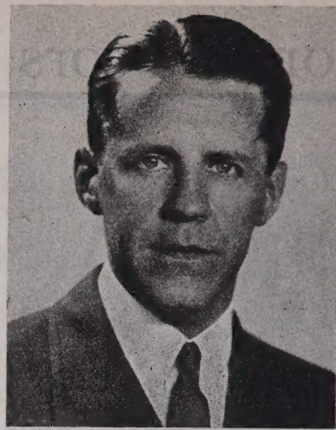
From 1924 to 1926, Mr. Herold was with the Bell Telephone Laboratories, and from 1927 to 1929 with E. T. Cunningham, Inc. In 1930 he entered the research and engineering department of the RCA Manufacturing Company, at Harrison, N. J. Since 1942, he has been associated with the RCA Laboratories at Princeton, N. J. He is a member of Phi Beta Kappa and Sigma Xi.

Rodney W. Johnson (S'42–A'45–M'45) was born in San Francisco in 1918. He received the B.S. degree in electrical engineering from the University of California. In 1940 and 1941 Mr. Johnson was with the Consolidated Aircraft Corporation and the Permanente Metals Corporation. From 1942 until 1945 he was with the Radiation Laboratory, University of California, as an electronic engineer. He has held amateur radio license W6MUR since 1935, and has been actively engaged in electronic research and development for several years. He has recently been engaged as an instructor in the Engineering Science Management War Training program, in addition to his work at the Radiation Laboratory. He is now operating his own company engaging in engineering, maintenance and installation of electronic equipment on the west coast. He is a member of Eta Kappa Nu and an Associate of Sigma Xi.

W. Rupert Maclaurin was born in Wellington, New Zealand. He received the A.B. degree from Harvard College in 1929, the M.B.A. degree in 1932, and the Ph.D. degree in 1936, from the Harvard School of Business Administration.

From 1933 to 1936 he was research assistant, traveling fellow, and instructor at the Harvard School of Business Administration. He served at the Buenos Aires branch of the First National Bank of Boston, in 1931; as statistician for the Boston Safe Deposit and Trust Company, from 1932 to 1933, and has been consultant for a number of industrial companies. He is at present connected with the Sylvania Electric Products Company, and serving as a special adviser on postwar plans for science and engineering, in the Office of Scientific Research and Development.

Dr. Maclaurin is professor of economics at the Massachusetts Institute of Technology, and has been director of the Industrial Relations Section there since 1937.



Royal Atelier

W. Rupert Maclaurin

Lawrence A. Ware (A'41) was born at Bonaparte, Iowa, on May 21, 1901. He received the B.E. degree in electrical engineering in 1926, the M.S. degree in physics in 1927, the Ph.D. degree in physics in 1930, and the E.E. degree in 1935, all from the University of Iowa. From 1929 to 1933 Dr. Ware was a transmission engineer with the Bell Telephone Laboratories; from 1935 to 1937 he was assistant professor of physics at Montana State College; and since 1937 he has been assistant and associate professor of electrical engineering at the State University of Iowa. He is a member of the American Institute of Electrical Engineers, the American Physical Society, Society for the Promotion of Engineering Education, American Association of University Professors, Eta Kappa Nu, Sigma Xi, and Tau Beta Pi.

For a biographical sketch of Chandler Stewart, Jr., see the January, 1945, issue of the Proceedings.

Lawrence A. Ware